TRANSISTORS

THEORY *and* PRACTICE

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SECOND EDITION
REVISED and ENLARGED

D. B. TARAPOREVALA SONS & CO. PVT. LTI
TREASURE HOUSE OF BOOKS
210 Dr. DADABHAI NAOROJI ROAD, BOMBAY 1.

TO MARY

First Printing–December, 1958
Second Printing–February, 1959
Third Printing–May, 1959
Fourth Printing–March, 1960
Fifth Printing–September, 1961
Sixth Printing–July, 1963

Second Indian Reprint 1966
Third Indian Reprint 1969

This book has been published with the assista
Joint Indian-American Textbook Programme.

It includes every word and illustration contai
latest original higher priced edition of this titl
complete and authentic copy thereof.

Sales Territory : All countries of Asia and Africa.

This edition can be exported from India only by the
D. B. Taraporevala Sons & Company Private Limi

Price in India: Rs. 3/-

Printed and published by Russi J. Taraporevala for D. B.
& Co. Private Ltd., 210, Dr. Dadabhai Naoroji Road,
Electregraphic Industries Division, Apte Industrial Estate,

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Foreword to the Second Edition

Since 1954 when this book was first published, the transistor has been widely accepted and greatly improved. Regarded at first as a somewhat fragile and unreliable contender for the position monopolized by the vacuum tube, the transistor now has, by virtue of its proven ruggedness, reliability and long life, taken over some of the functions of the tube, especially in miniaturized equipment.

In attempting to bring this book up to date, we have corrected the statements regarding transistor limitations which were acceptable at the time of original publication, made additions and changes consistent with improvement of the art, removed obsolete material, and expanded the bibliographies. Although at present the point-contact transistor has been superseded largely by junction types, we have not deleted references to the theory and applications of this earlier type except where specific models no longer manufactured were mentioned, since improved methods of manufacture and quality control might possibly revive the point-contact unit for specific applications.

New transistors appear regularly and the ratings of older ones are as regularly improved. Therefore, printed tables rapidly grow obsolete. For this reason, we eliminated the chapter which originally gave the ratings and sizes of commercial transistors and have devoted this space to expansion of the other material in the book.

In a field which advances as rapidly as that of semiconductors, obsolescence takes place almost faster than printer's ink can dry. For a useful book to keep abreast of the art, periodic revisions are a necessity. It is hoped that the present revision will render this edition of *Transistors, Theory & Practice* as acceptable to teachers, students and technicians as its predecessor was

introduction

THE purpose of this book is to provide an elementary explanation of transistor theory and operation for the thousands of practical electronic workers.

A comparatively enormous quantity of periodical literature appeared during the transistor's early years. Unfortunately, however, much of the explanatory material, chiefly from the pens of advanced physicists, has been so involved mathematically as to be completely useless to the practical man and not too well understood by some engineers.

In this book, I have tried to tell in simple language how transistors work and what their circuits are like. I hope I have achieved my purpose without too often falling into the sin of oversimplification. It is hoped that this work will fill the present need for an introductory text.

Assisting in the effort have been numerous organizations which have furnished data or illustrations. Grateful acknowledgement is made here of such material, and for the permission to make use of it, received from Bell Telephone Laboratories, CBS-Hytron, Federated Semi-Conductor Co., General Electric Co., Hydro-Aire, Inc., National Bureau of Standards, Radio Corporation of America, Radio Receptor Co., Inc., Raytheon Manufacturing Co., Sylvania Electric Products, Inc., Texas Instruments, Inc., Transistor Products, Inc., Western Electric Co., and Westinghouse Electric Corp.

Data on the surface-barrier transistor were obtained through the courtesy of the Philco Corporation and Proceedings of the IRE.

A tremendous amount of work, absorbing some of the most talented minds in the field, is constantly being carried on in transistor electronics. Undoubtedly, this will result in many new theories, reworked older ones, further transistor types and new circuits. In the meantime, the text is offered with the sincere hope that it will enable many practical men to take their first step less falteringly into the realm of the transistor.

semiconductor theory

wafers, used in diodes and transistors, often are called crystals for the same reason.

Inside the crystal lattice, certain loosely bound electrons (called *valence electrons*) in the outer rings of one atom align themselves with similar electrons in adjacent atoms to form *valence bonds* which hold the atoms together in the orderly structure of the lattice. Thus, in any valence bond there are shared electrons, so called because they are shared by neighboring atoms.

Electrons and holes

Some crystalline materials have electrical characteristics which may be regarded as intermediate between those of conductors and those of insulators, hence are termed *semiconductors.* Under ordinary conditions, semiconductors are neither good insulators nor good conductors, but can be made to exhibit some of the properties of each. Among the semiconductors having practical importance in modern electronics are cadmium sulphide, copper oxide, copper sulphide, germanium, lead sulphide, selenium, silicon, and silicon carbide (carborundum). Many other elements and compounds have been found to possess semiconductivity in varying amounts.

To understand the nature of a semiconductor, it is necessary to look into the atomic arrangement of the crystal. At low temperatures, no electrons in a true semiconductor are available to carry current through the material because the loosely bound electrons which ordinarily would be available for that purpose are held in the valence bonds. The material therefore has the features of an insulator, or at least of a very-high-value resistor. If the crystal structure were perfect and all valence bonds satisfied, the material would support no current flow at all and would be an insulator in the truest sense. Based upon advanced quantum-mechanical theories of matter and verified by experiments, we know that a crystal of *pure* germanium (one having identical atoms uniformly spaced) will not conduct at all. The germanium diode, widely used as a detector in radio and TV, conducts because of the presence of impurities. The atoms of a crystal of germanium, like all other atoms, contain a nucleus surrounded by rings of electrons. These rings, often likened to a miniature solar system with the sun analogous to the nucleus, remain *bonded* to the nucleus. Electrons, however, can be added to or taken away from the outermost ring. Atoms of the element phosphorus have an outer ring which contains one more electron than the outer ring of germanium. Adding

an *excess* electron condition. *Particularly at higher temperatures,* thermal agitation causes some of the valence electrons to be knocked out of the bonds and thus to become available for current flow. The material then assumes the features of a conductor. As the semiconductor material is heated further, more electrons are freed to drift through the crystal lattice in response to an applied electric field, and the conductivity of the material increases (resistance decreases). Quite apart from heat action, electrons in some semiconductors may also be dislodged from the valence bonds at ordinary temperatures by the action of light shining upon the material. This photoelectric action is utilized in the selenium photocell of the photographic exposure meter. Bombardment by other forms of radiation likewise has been observed to release electrons in semiconductors.

A somewhat different situation exists when we add a material which contains *less* electrons in its outermost orbit when compared to germanium. Elements such as boron, indium, aluminum, or gallium, have *one less* electron than germanium. Since we now have an inadequate number of electrons surrounding the nuclei of the atoms, we say that the orbit has *holes* in it. Furthermore, these holes can move just as though they were positive charges. This apparently fantastic concept can be mathematically proved and verified experimentally. However, you can consider it in this manner. An electron is a basic negative charge. When an electron jumps out of a valence bond, it leaves behind an empty *hole* which, representing a deficiency of negative charge, appears as a net positive charge. Another electron, under attraction by the positive electrification, may jump into the hole. This leaves a hole farther back which then may be filled by a nearby electron, leaving still another hole. It is in this way that holes are conceived of as drifting through the lattice. Usually, a certain amount of recombination goes on between holes and electrons, a hole–electron pair recombining to restore the original condition of the material.

Some imperfect semiconductor materials have an excess of electrons. In these materials, current flow results principally from electron movement. Such semiconductors are termed *n-type* (*n* signifying the negative polarity of the electron). Other semiconductors have an abundance of holes. Conduction in these latter materials is by the drift of holes. Hole-rich semiconductors are termed *p-type* (*p* referring to the positive polarity of the hole). Thus, the type of current carrier determines the classification of the semiconductor material. *Intrinsic* semiconductors are those materials in which

only a small amount of energy is required to displace electrons from valence bonds. At ordinary temperatures, electrons and hole are generated in pairs and recombine continuously in intrinsi semiconductors. The useful intrinsic semiconductor elements fo electronic applications are in the 4th column of the Periodic Table The Periodic Table is found in most elementary textbooks o chemistry.

When an electric field is applied to a piece of intrinsic semicon ductor material (for example, by placing a positive electrode o one end and a negative electrode on the other), electrons drif through the crystal lattice toward the positive electrode and hole toward the negative electrode. This constitutes a flow of curren Electron movement and hole movement may be regarded as sepa rate manifestations of the same phenomenon. Although electron and holes move simultaneously in opposite directions, the curren component which their separate movement constitutes is in on direction.

The conduction characteristics of a semiconductor can be altere considerably by mixing a minute quantity of a selected impurity metallic or nonmetallic, into the semiconductor material. Atom of the impurity material must be small enough to fit into th crystal lattice where they replace a few of the semiconductor atom When the atoms of the impurity material have more valence ele trons than are required to satisfy the valence bonds with adjacen semiconductor atoms, loose electrons will be left over and will b free to participate in current conduction. Such an impurity there fore will make the semiconductor strongly n type. An impurity c this type is termed a *donor* since it donates electrons. If, insteac the impurity atoms have less valence electrons than are needed t satisfy the valence bonds, holes will occur in each bond where a electron is missing. These holes are available for current condu tion and the semiconductor material will be strongly p type. A impurity of this kind is termed an *acceptor*, since the holes creates accept electrons.

Thus, by proper choice of the impurity material with respect t its own electronic constitution, the semiconductor may be mac either n type or p type as desired. It is interesting to note that th proper amount of added impurity in some transistor semicondu tors may be as small as 1 impurity to each 10 million semiconducto atoms. When the original semiconductor material is an elemer from column 4 of the Periodic Table, donor impurities can com from column 5 and acceptor impurities from column 3.

Fig. 101 is a simplified representation of a donor-impurity atom in part of a semiconductor crystal lattice. Here, each semiconductor atom has four valence electrons and the impurity atom has five. Four of the valence electrons of the impurity atom join with those of adjacent semiconductor atoms to form valence bonds. All

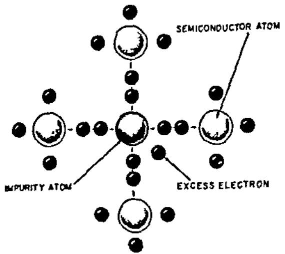


Fig. 101. Donor-impurity atom in a semiconductor.

valence bonds being satisfied, the extra electron of the impurity atom is not held firmly in the structure and is free to move through the crystal lattice. Mixing atoms of this type into the semiconductor thus makes the latter n-type.

Fig. 102 is a simplified picture of an acceptor-impurity atom in part of a semiconductor crystal lattice. In this case, as before, each semiconductor atom has four valence electrons. But the impurity has only three. Only three valence bonds with neighboring semiconductor atoms therefore are satisfied. (A fourth impurity electron would be required to complete the bond between the impurity atom and the top semiconductor atom.) A hole accordingly appears in what would have been the upper bond. An available electron from somewhere else in the lattice can move into this particular hole, thus leaving a hole behind. The hole provided by the impurity atom thus can migrate through the lattice under the proper conditions. Mixing this type of impurity into the semiconductor makes the latter p-type.

It must be remembered that these are purely graphic representations of hypothetical semiconductors and impurities in which all of the inner electrons and the nuclei of each of the atoms have

been ignored. The electrons do not necessarily align themselves in the simple manner shown. Nor has any attention been paid to orbital movement of the electrons. However, the pictures presented give some background for understanding how semiconductor properties are modified by doping these materials with selected impurities.

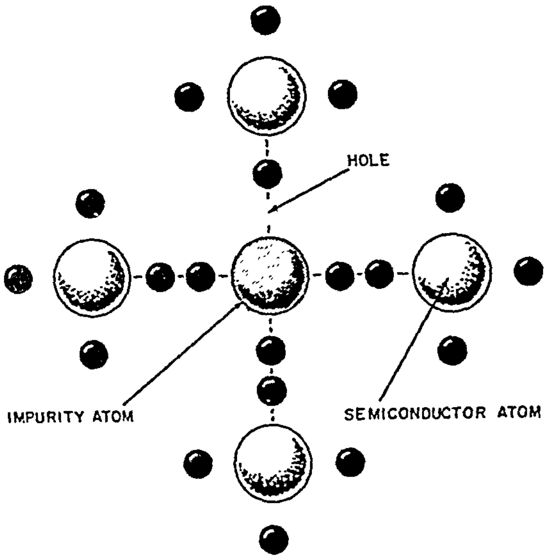


Fig. 102. *Acceptor-impurity atom in a semiconductor.*

It has been mentioned that electron and hole activity within the semiconductor crystal lattice is accelerated at high temperatures. Conductivity increases during heating. Semiconductors therefore exhibit a negative temperature coefficient of resistance.

Current moves more slowly through a semiconductor than in a true conductor. Electrons drift more slowly because they encounter obstructions due to crystal imperfections. Hole movement is even slower because of their jump-by-jump progress between valence bonds. The total current may fall short of expected values because of the devious routes taken by some of the carriers, a phenomenon known as *spreading*.

Germanium

At this writing, germanium and silicon are the most extensively used semiconductors in the manufacture of transistors. Silicon is important because of its ability to operate at higher temperatures.

Germanium is a grayish-white metallic element. In some of its properties it resembles carbon, while in others it resembles tin. It is found in the 4th column of the Periodic Table and has an atomic weight of 72.60.

Germanium was predicted in 1871, although never seen, by Dmitri Mendelyeev who called it eka-silicon. It was discovered physically by Clemens Winkler in 1886 and was named in honor of Germany, Winkler's native country.

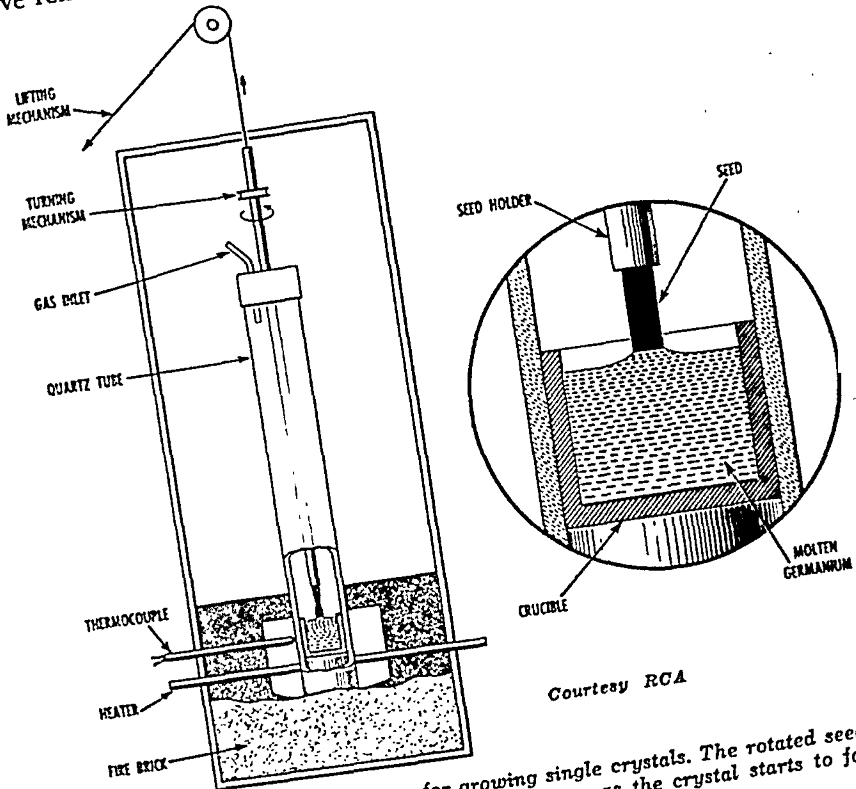
Metallic germanium is secured by reduction (in a hydrogen or helium atmosphere) of germanium dioxide, a gray powder. The dioxide is obtained in commercial quantities in the United States as a flue residue in zinc smelting, and in England as a component of the chimney soot from gas works.

After purified germanium has been produced by electronic manufacturers and doped to specifications, comparatively large single crystals of it (as shown in Fig. 103) are drawn from a melted mass of the metal by dipping in a seed crystal of germanium and withdrawing it slowly under rotation. The melted germanium adheres and follows the seed to be pulled out of the melt in single-crystal form. During the process, temperature is controlled closely and air is excluded. The tiny germanium wafers used in diodes and transistors later are cut out of this single crystal. The advantages of single-crystal material are uniformity and reproducible electrical characteristics, such as resistivity. When, on the contrary, a germanium sample is composed of numerous intimately bonded separate crystals, wafers sliced from this material might cut through crystal interfaces and exhibit nonuniformity of characterstics due to separate crystal properties.

While the germanium is in the molten state, impurities of the proper kind and amount are added to make it either n-type or p-type, as required. Without controlled doping, pure germanium would behave like an insulator. Later, during the single-crystal drawing, impurities may be added at proper times during the withdrawal to produce separate n and p layers in the same crystal. Most general-purpose germanium is prepared to be n-type.

Silicon, like another well-known semiconductor material selenium, is a non-metallic element. Like germanium, silicon is found in the 4th column of the Periodic Table and behaves similar to carbon in many chemical reactions. It has an atomic weight of 28.06. Although silicon is abundant in the earth's crust, being second only to oxygen, it never occurs in nature in the free state. Many natural silicon compounds are complex. Because of

this fact, silicon-bearing materials must be put through an extensive refining process to obtain silicon of the high purity required



Courtesy RCA

Fig. 103. *One type of apparatus for growing single crystals. The rotated seed i[s] slowly withdrawn. The insert shows early stage as the crystal starts to form*

by semiconductor devices. After purification, silicon is doped wit[h] a controlled amount of a suitable impurity element to give useful semiconductor properties.

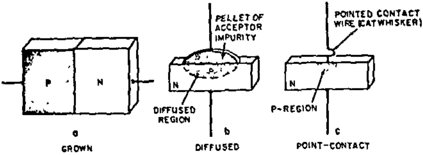
P-n junction, semiconductor diode

The junction between a p region and n region in single-cry[stal] germanium is of interest, since it is the foundation of both di[ode] and transistor action.

Fig. 104 shows three types of such junctions. In Fig. 104 and n regions have been *grown* into the germanium bloc[k] mixing acceptor and donor impurities, respectively, into th[e] gle crystal during its formation. This is known as a *grown*

tion. Note that the grown type of p-n junction is not a sandwich made by attaching a p block to an n block, but actually consists of p and n layers in a single piece of germanium. Considerable misunderstanding has arisen regarding this arrangement.

The *diffused junction* in Fig. 104-b is made by placing a pellet of acceptor impurity, such as indium, on one face of a wafer of



Figs. 104-a, -b, -c. *Types of p-n junctions (diodes).*

n-type germanium and then heating the combination to melt the impurity. Under proper conditions of temperature and time, a portion of the impurity metal will diffuse a short distance into the wafer, thereby creating a region of p-type germanium in intimate association with n-type bulk. This is also called an *alloyed junction* or a *fusion-alloy junction* from the fact that a small amount of the pellet material alloys with the germanium.

Fig. 104-c shows a point-contact type. Here, a fine, pointed wire ("catwhisker") makes pressure contact with the face of an n-type germanium wafer. After assembly, the device is electroformed by passing a high-current surge momentarily across the junction of wafer and whisker. The heat generated during the short interval drives a few electrons from the atoms in the region of the point contact, leaving holes and thus converting into p-type a small volume of germanium immediately under and around the point.

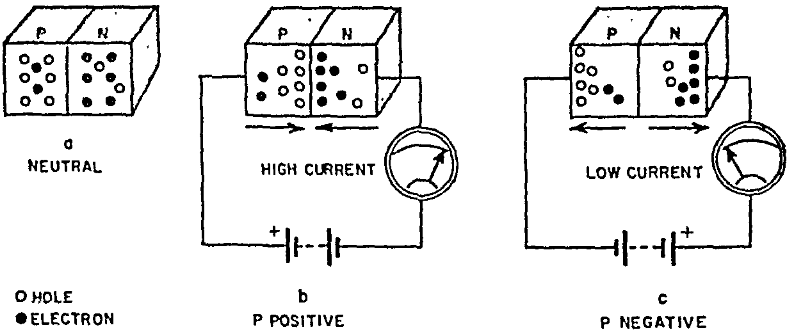
Silicon p-n junctions are produced in a similar manner. In most instances, the silicon has been processed in such a way as to make it p-type. To create the junction, an n-type material is either inserted at the proper point in the crystal growing process or (in the diffused-junction process) an n-type material is later diffused into the body of p-type silicon wafer Like germanium diodes, silicon diodes also are produced both in the junction and point-contact types.

Fig 105 shows what happens when a steady dc bias voltage is applied between the p and n portions of a junction. In Fig. 105-a,

no voltage is applied. The p region has holes (white), and the n region electrons (black) available as carriers. Note that a few electrons may be seen in the p region, and a few holes in the n region. But in each case, these *minority carriers* are far outnumbered by the *majority carriers* and can make no substantial contribution to any current conduction.

When the p region of the junction is made positive, as in Fig. 105-b, the holes are repelled by the positive field and the electrons by the negative field. Both holes and electrons are driven, in the direction of the arrows, toward the p-n junction where they recombine. A high current flows, since the junction resistance appears low. The process continues as long as the bias voltage is applied.

When the p region is made negative, as in Fig. 104-c, holes are attracted by the negative field and electrons by the positive field. Holes and electrons both are pulled away from the p-n junction, in the direction of the arrows. There can be no significant recombination, and the junction resistance appears high. As a result of this action, the current flow is low.



Figs. 105-a, -b, -c. *Effect of junction bias voltage.*

We say that the junction is biased in the *forward direction* in Fig. 105-b. This is the direction of low resistance, high conductance, or high current. Conversely, the junction is biased in the *reverse direction* in Fig. 105-c. This is the direction of high resistance, low conductance, or low current. A *potential barrier* is said to be set up when the junction is reverse-biased.

The p-n junction is a rectifier because of this ability to pass current more readily in one direction than in the other. Its rectification efficiency is proportional to the ratio of its forward and reverse resistances. This is the diode rectifier which has been the most

widely exploited semiconductor device. Two varieties of diodes are manufactured—the junction type (of which Figs. 104-a and b are illustrative) and the point-contact type, the basic arrangement of which is shown in Fig. 104-c. However, if the concept of a formed p-area around the point-contact whisker is accepted, then all diodes may be regarded as being of the junction type.

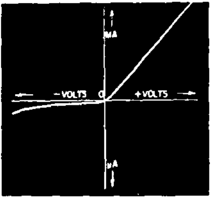


Fig. 106. Static volt-ampere characteristic curve of a typical semiconductor diode.

Fig. 106 shows the static volt-ampere characteristic curve of a typical semiconductor diode. Note that the forward current is in milliamperes, while the reverse current is only a few microamperes. The positive and negative portions of the curve are seen to be nonlinear over a considerable part of their ranges.

Fig. 107 shows the static resistance (E/I) characteristic of a semiconductor diode. As the forward voltage is increased, the resistance

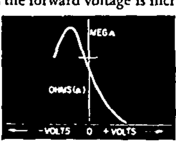


Fig. 107. Static resistance characteristic of a semiconductor diode.

falls to a low value, usually 100 ohms or less. At decreasing values of forward voltage, the resistance increases until, near zero voltage, it is in the hundreds of thousands of ohms. As the reverse volt-

age is increased, the resistance passes through a peak in the hundreds of thousands of ohms (or in the megohms) and then decreases. There is a burnout point at the positive and negative extremes of this curve, as in the curve in Fig. 106.

Hole and electron injection

When a positive bias is applied to the p region of a p-n junction or to the whisker of a conventional point-contact germanium diode, valence electrons from nearby atoms flow to the p region or to the whisker. This is equivalent to stating that holes flow *from* the p region or whisker. Another point of view is that, with the same bias polarity, the n-region injects electrons into the p-type germanium.

This concept of carriers (whether holes or electrons) being injected into the body of the semiconductor material is essential to an understanding of the mechanics of transistor operation.

In the next chapter, we shall see that carrier injection into a semiconductor is a basic phenomenon which is analogous to electron emission by the cathode of a vacuum tube.

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transistor characteristics

THE transistor is a semiconductor amplifier device. Its amplifying and control properties suit it also to oscillator and switching functions. In operation and application, the transistor bears somewhat of a resemblance to the vacuum tube, but is different from it in several important respects In the tube, electrons are liberated into an evacuated space, and their drift toward a positively charged plate electrode constitutes a current flow. Amplification results from the fact that this current can be controlled and modulated by a small signal voltage.

In the transistor, either holes or electrons are injected into the solid body of a semiconductor and their movement through the material constitutes a current flow which likewise can be modulated by a signal voltage. The fundamental differences between tubes and transistors result from the mechanism of control. In the tube, the electron current is modulated electrostatically by a signal *voltage*. Under ordinary circumstances, no signal current is required. The tube thus is a voltage-actuated high-impedance device. But in the transistor, signal energy is required to modulate the injection of carriers into the semiconductor to modulate the carrier current. This corresponds to current variation. A signal source employed with a transistor accordingly must deliver current, and the transistor is a low-impedance current-actuated device.

Transistor types

Most present practical transistors are triodes (although tetrodes are becoming increasingly important in high-frequency applica-

tions), having three electrodes which correspond roughly to cathode, grid, and plate of a triode tube. The two types of triode are

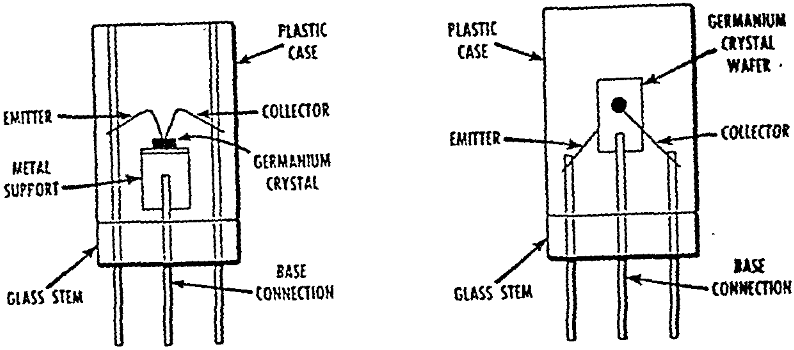
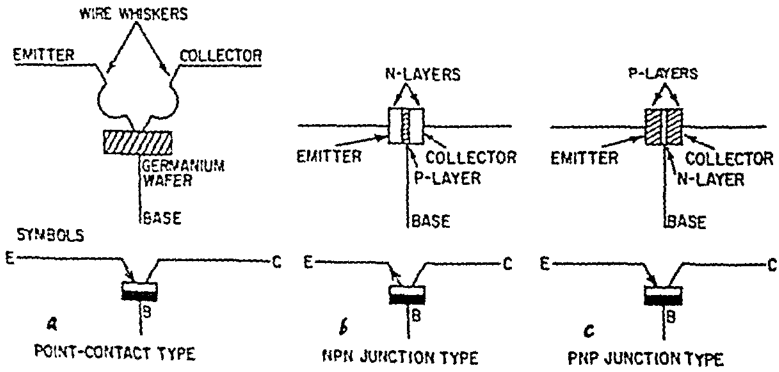


Fig. 201. *Essential features of point-contact and junction transistors.*

the *point-contact* and *junction* transistors. Fig. 201 shows their main features.

Each basic type of triode transistor has the following electrodes: an *emitter* which serves to inject carriers into the *base,* and a *collector* which attracts the carriers through the base-region. The emitter corresponds to the cathode of a vacuum tube, the base to the grid, and the collector to the plate.

Fig. 202-a shows the arrangement of a point-contact transistor. This type resembles a point-contact diode with an extra cat-



Figs. 202-a, -b, -c. *Point-contact and junction type transistor triodes.*

whisker. The two whiskers make pressure contact with the face of a germanium wafer (commonly n-type). One whisker serves

as the emitter, the other whisker as the collector, and the germanium wafer as the base. The points of the whiskers are closely spaced on the germanium, usually being separated by .002 or .003 inch. A large-area, low-resistance contact is made to the base. The three connections terminate in pigtails or pins for external access.

Fig. 202-b shows an ideal cross section of a junction transistor of the n-p-n type. Here, careful processing has produced a thin p-type layer between two n-type layers in the same block or wafer of single-crystal germanium. One n-layer is the emitter, the other n-layer the collector, and the central p-layer is the base. The latter is very thin, often .001 inch. Low-resistance connections to the three layers terminate in pigtails or pins.

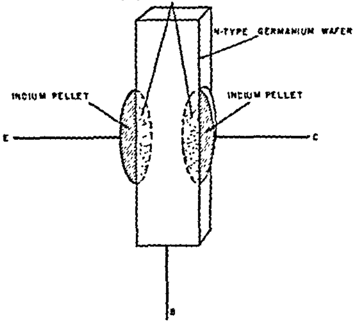


Fig. 203. *Construction details of a diffused p-n-p transistor*

Fig. 202-c illustrates an ideal cross section of a junction transistor of the p-n-p type. In this unit, a thin n-layer is processed between two p-type layers in the same block or wafer of germanium. Here, the p-layers are emitter and collector, and the central n-layer is the base. The n-layer is very thin, often .001 inch or less. Low-resistance connections to the three layers terminate in pigtails or pins.

The point-contact transistor is seen to resemble the point-contact diode (Fig. 104-c), and the junction transistors the junction diode (Fig. 104-a and Fig. 104-b), with one extra electrode provided in each instance. In Fig. 202, the corresponding standard symbol is shown with each type of transistor. Note that the same symbol is used for both point-contact and p-n-p types.

As in the junction diodes, the layers of the junction transistors may be of the grown or diffused type. The grown type is obtained by adding the required impurities during the process of single-crystal pulling to create the adjacent n and p layers. This is the most difficult type to manufacture. Fig. 203 shows the cross-section

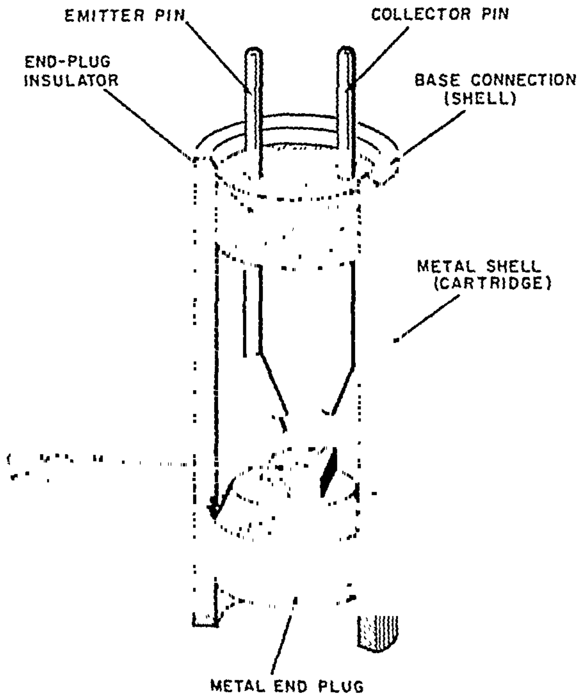
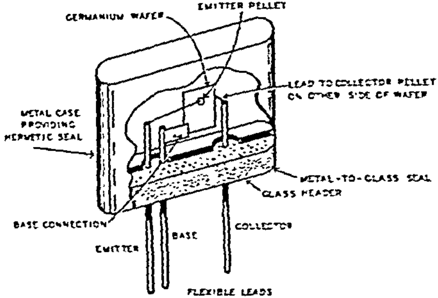


Fig. 204. Cutaway view of a point-contact transistor assembly.

of a diffused-junction p-n-p type. Here, a pellet or button of indium is melted on each face of a thin wafer of n-type single-crystal germanium. Some of the indium diffuses into the germanium from each side, creating p-type regions in the wafer. The process is continued until the separation between the two p-regions is very narrow (say .001 inch) but is halted before short circuit

occurs. In practical diffused-junction p-n-p transistors, the emitter pellet is somewhat smaller in size than the collector pellet.

Transistors of both types are manufactured to several mechanical designs. Fig. 204 is a cutaway view of one style of point-contact transistor. This type is housed in a metallic cartridge, shell, or barrel and is intended for insertion into a special subminiature socket. The two pins fit into the socket clips, while the transistor shell (base connection) is gripped by the socket ring or spring. The germanium wafer is soldered to the metal end plug or to a



Courtesy CBS Hytron

Fig. 205 Cutaway view showing inner construction details of a junction transistor.

metal pin passed through this plug. The plug makes a tight-fit contact with the shell to form the low-resistance base connection.

Fig. 205 is a cutaway view of a junction transistor of one style of construction. Here, the germanium wafer is supported vertically. A large metal tab is attached to the lower end of the wafer to form a low-resistance base connection. The opposite end of the tab is attached to the base lead or pin. The emitter, base, and collector leads pass through the insulated bottom of the assembly and are intended for insertion into the clips of a subminiature tube socket or for soldering or welding into a circuit. The assembly shown in Fig. 205 is used for hermetically sealed and evacuated transistors as well. After assembly, a protective wax is sometimes

injected into nonevacuated transistors through a hole provided for the purpose, after which the hole is sealed.

By comparison with point-contact and junction diodes, the emitter-to-base section of the transistor forms a diode, and the collector-to-base section a second diode. The base region is a common electrode for both diodes.

Surface-barrier transistor

A third class of transistor, the *surface-barrier* type, has been developed and extensively tested by the Philco Corp. While it resembles the n-p-n junction type somewhat in general appearance, the surface-barrier transistor is unique in that it contains

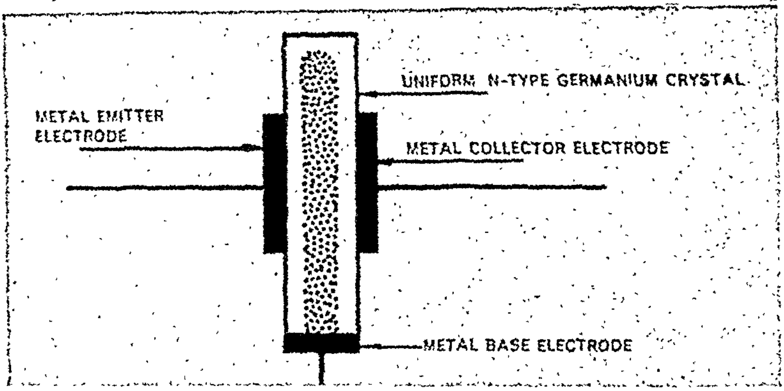


Fig. 206. *Schematic cross-section of a surface-barrier transistor.*

only one type of germanium, n-type. Metal emitter and collector electrodes of indium are *plated* on opposite faces of an extremely thin portion of an n-type germanium wafer cut from single-crystal material. See Fig. 206. No diffusion of the metal takes place into the germanium.

Beneath the emitter and collector electrodes lies a thin barrier layer, approximately 1/10,000-inch thick, which contains almost no electrons or holes. This is represented by the white region in Fig. 206. The barrier layer extends slightly into the crystal. It is an almost perfect insulator and has a strong electric field and so is analogous to a capacitor. The field repels free electrons *into* the crystal, from the surface. The free electrons from the interior atoms of the crystal are in effect driven back down and the barrier remains swept clean of such carriers.

Surface-barrier transistors are produced by an electrolytic etching and plating process: Two fine streams of indium sulphate solu-

tion are played upon axially opposite points on the faces of the n-type germanium wafer. At the same time, a direct current is passed through the germanium and solution in such a direction as to remove germanium electrolytically from the faces of the wafer. The tiny sprayed areas thus are gradually etched away. When the desired wafer thickness (a few ten-thousandths of an inch) has been reached, the etching process is abruptly arrested by reversing the direction of current flow. This reversal causes an indium-metal dot to be plated on each opposite face of the etched-out area. See Fig. 207. Leads then are attached to the emitter and

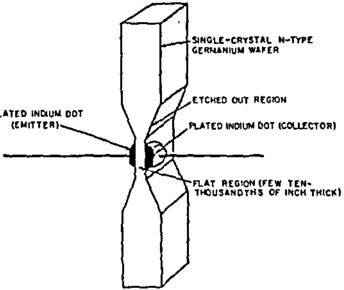


Fig. 207. *Detail of the cross-section of a surface-barrier transistor.*

collector dots and to the wafer (base), and the unit is mounted in a hermetically sealed enclosure similar to Fig 205.

The electrolytic process has several advantages: (1) The wafer thickness can be controlled within a few millionths of an inch; (2) the opposite faces of the etched-out area are almost perfectly flat and parallel; (3) the liquid streams keep the work both clean and cool; and (4) the plating process is effected on a clean germanium surface *immediately* after the etching, precluding any possibility of contamination.

Advantageously small-sized electrodes are obtained with the electrolytic etching and plating process. Philco reports typical

units with .003-inch emitters and .006-inch collectors and a barrier spacing of .0002 inch.

High-frequency junction transistor

Many techniques have been devised for producing high-frequency junction transistors. One such triode transistor is illustrated in Fig. 208. In this type, a small area of the single-crystal semiconductor wafer is drilled almost through, leaving a very thin wall. A tiny dot of p-type alloying metal (such as indium) is placed on each side of this wall and the emitter and collector junctions produced in the manner explained earlier for conventional diffused-junction transistors.

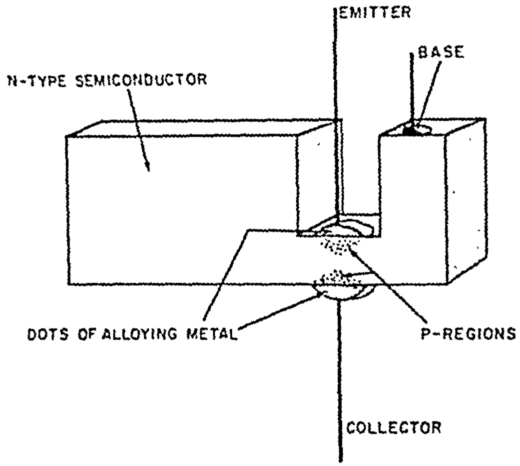


Fig. 208. High-frequency junction-transistor cross-section.

The improved electrical characteristics resulting from this type of construction permit operation of the transistor at frequencies higher than 10 mc. Additional characteristics are discussed later in this chapter.

Diffused-base transistor

In the diffused-base transistor (Fig. 209), the concentration of impurity centered in the very thin base zone is much greater near the emitter than near the collector. The shading of the base zone in Fig. 209 attempts to show this gradient. A "built-in" electric field results from the fact that these impurities effectively are ionized. This field accelerates any carriers injected by the emitter across the base zone and thereby improves operation at high fre-

quencies. Laboratory-produced germanium diffused-base transistors have been operated to frequencies between 500 and 600 mc. Both germanium p-n-p and silicon n-p-n transistors of this type have been produced.

In one method of making a diffused-base transistor, an n-type impurity (such as arsenic-doped germanium) is diffused into a p-type, single-crystal germanium bar.

Further characteristics of the diffused-base transistor are discussed later in this chapter.

Transistor action

The exact, detailed behavior of a transistor is described in the complicated mathematical language of solid state physics and

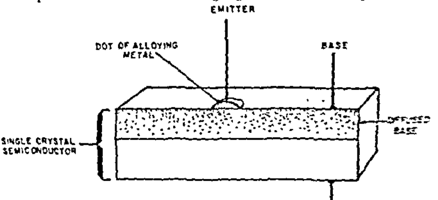


Fig 209, Cross-section of diffused-base transistor.

quantum mechanics. Abundant literature is available for those readers who are inclined toward advanced mathematics.

A first explanation of the basic mechanism of transistor operation can now be made with the aid of 10 condensed facts which we are in a better position to understand since we have looked into the subjects of diode operation and carrier injection.

1. A steady dc bias voltage (v_c) is applied between collector and base with polarity such that reverse current is forced through the collector diode.

2. The reverse-biased collector diode appears as a high resistance (r_c).

3. A steady dc bias voltage (v_e) is also applied between emitter and base with polarity such that forward current flows through the emitter diode.

4. The forward-biased emitter diode appears as a low resistance (r_e).

5. The emitter injects carriers into the base-region. The polarity of these carriers is opposite to the polarity of the collector. The carriers accordingly are drawn through the base-region toward the collector by action of the latter's field.

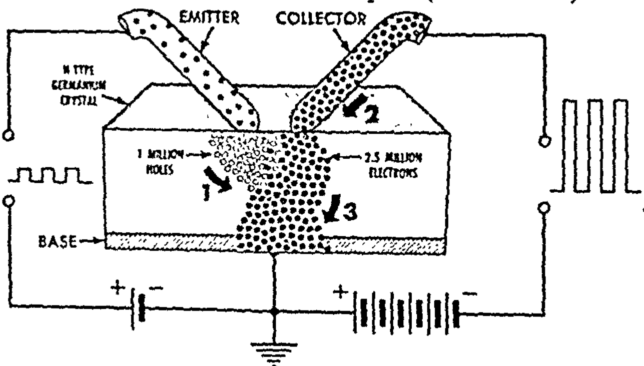
6. On their way to the collector, a few of the carriers recombine with carriers of opposite polarity in the base region and thus are neutralized. But most of them reach the collector where they increase the collector current (i_c). This is the fundamental mechanism whereby current amplification (or at least emitter–collector control interaction) occurs in a transistor.

7. Current amplification may be expressed as the ratio of the change in collector current (di_c) to a given change in emitter current (di_e) when collector voltage (v_c) is held constant. *Current amplification factor* (di_c/di_e) is represented by the Greek letter alpha (α).

8. When every carrier leaving the emitter reaches the collector, alpha equals 1. In other words, the collector-current change is equal to the emitter-current change. Alpha is less than 1 (but close to unity) in junction transistors, and can be greater than 1 in point-contact transistors.

9. The transistor exhibits voltage gain and power gain, as well as current gain when it is operated in satisfactory circuits.

10. Voltage gain and power gain are obtained even when alpha is less than 1, because the output (or collector) electrode

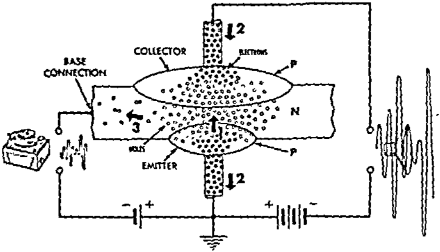


Courtesy RCA

Fig. 210-a. Enlarged point-contact transistor: If a signal injects 1 million holes at the emitter, they will be attracted toward the collector (1). Near the collector, the holes reduce the barrier to electron flow (2) allowing some 2.5 million electrons to pass into the crystal. Of these, 1 million neutralize the holes; the others flow to the base (3). Pulses at the left and right are of the type employed in computers.

has a higher impedance than the input (or emitter) electrode. However, the emitter is not always used as the input electrode, nor the collector always as the output electrode.

Figs. 210 a, -b are enlarged sketches showing how bias and signal voltages are applied to transistors. Circuit diagrams illustrating transistor connections are shown in Fig. 211. In the n-p-n junction transistor (Fig. 211-a) the current carriers are electrons. The negatively biased emitter injects electrons into the base region. Under the influence of the strong positive field of the collector, these electrons diffuse through the thin base region to the collector junction where they augment the collector current. A few of the electrons combine with holes in the hole-rich p-germanium base



Courtesy RCA

Fig. 210-b. Enlarged junction transistor: A small signal from a phonograph is amplified to activate the speaker. If the signal changes by 1 million electrons, for example, there will be a voltage difference between the emitter and base which starts 50 million holes flowing out of the emitter (1). All but 1 million holes get to the collector, inducing 49 million electrons to flow and carry current in the collector circuit (2). The remaining holes flow to the base completing the base-emitter circuit (3).

region. But the base layer is very thin, so most of the electrons reach the collector. However, since recombination and other effects prevent *all* the electrons from reaching the collector, alpha (current amplification factor) for the n-p-n transistor can never reach a value of 1, although it attains values of 0.98, 0.99 and higher in some commercial units. The emitter bias is negative for the n-p-n type and the collector bias is positive. This is similar to negative grid bias and positive plate voltage in vacuum-tube

ircuits. The input signal is applied in series with the emitter bias n the example shown, and the amplified output signal is developed across the collector load resistance.

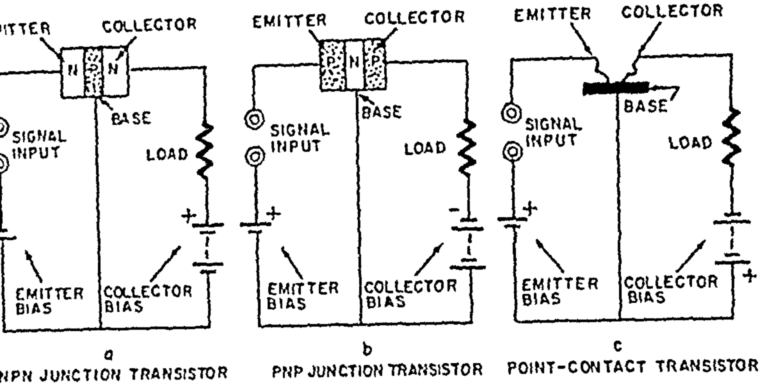


Fig. 211. Connections for junction and point-contact transistors.

In the p-n-p junction transistor (Fig. 211-b), the emitter bias s positive and the collector bias negative. The emitter injects positive holes into the n-type base region, and these holes are attracted by the strong negative field of the collector. A few holes combine with electrons in the electron-rich base layer and thus are neutralized. But the base layer is extremely thin and most of the holes diffuse through to the collector junction where they increase the collector current. As in the n-p-n junction transistor; not all of the holes can reach the collector, so alpha never quite equals 1, although it reaches values of 0.99 and higher in some commercial p-n-p units. The input signal is applied in series with the emitter bias in the example shown, and the amplified output signal is developed across collector load resistance.

Fig. 211-c shows connections to a point-contact transistor. Here, the bias-voltage polarities are the same as for the p-n-p junction transistor. The positively biased emitter injects holes into the n-type germanium base region, and these holes are attracted by the strong negative field of the nearby collector. Some recombination of holes and electrons takes place in the electron-rich base region, but most of the holes reach the collector where they augment the collector current. It is intriguing to note at this point that (unlike the current amplification factor of less than unity for n-p-n and p-n-p junction transistors) alphas higher than 1 are observed with the point-contact transistor. Values of 2 or 3 are

common in standard manufactured units, and much higher values have been recorded for experimental units.

There is considerable theory as to how high alphas are obtained with point-contact transistors. One would expect that an alpha of 1 would represent the condition of *all* injected carriers reaching the collector, and for this reason it would seem that alpha never could exceed unity. Several explanations have been offered. One is that a relatively strong positive field somehow associated with the *motion* of the injected holes permits each hole to accelerate more than one electron of collector current. Another (the "p-n hook theory"), more widely accepted at this writing, explains that the collector is a complex junction structure comprising a separate transistor all by itself and that this second transistor is an amplifier electron-coupled to a first transistor structure in the main body of the unit. This junction occurs as the result of electroforming the finished point-contact transistor.

In each type of transistor, the emitter bias voltage is lower than the collector bias voltage. This is understandable when we consider that the emitter diode is biased in the forward direction and consequently exhibits low resistance, while the collector diode is biased in the reverse direction and exhibits high resistance.

Transistor configurations

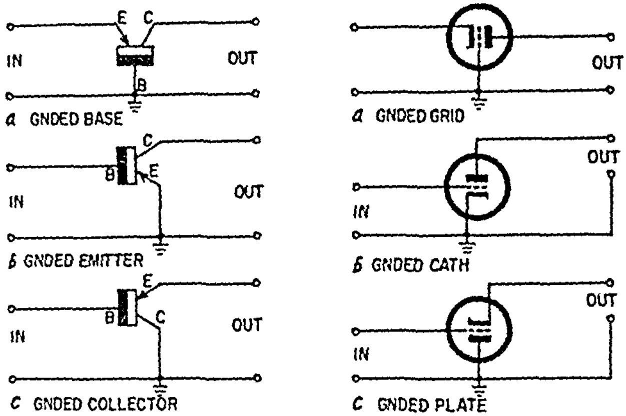
The emitter is not always used as the signal-input electrode, nor the collector always as the signal-output electrode. A transistor can be connected into a single-stage circuit in any one of three ways, depending upon the type of operation desired, and its amplifying or control characteristics depend upon which method of connection is used. It is advisable at this time to describe these three configurations to facilitate an understanding of the discussions which follow. This description is made here for purposes of identification *only*, and will *not* be complicated now by explanations of performance. They will come later in this chapter.

Fig. 212 shows the three basic configurations with approximately equivalent tube circuits for comparison. Transistor connections of the three types are employed in single-stage amplifiers, oscillators, trigger, and control circuits. Elaborations of all the basic configurations also are employed. *For simplicity, all bias-voltage supplies and external circuit components have been omitted.*

Fig. 212 a shows the *grounded base circuit*, also called *common-base*. The input signal is applied between emitter and base, and the amplified output signal is taken between collector and base.

This configuration is analogous to the grounded-grid tube circuit.

Fig. 212-b shows the *grounded-emitter* circuit, also called *common-emitter*. The input signal is applied between base and emit-



Figs. 212-a, -b, -c. *Transistor amplifier configurations with equivalent tube circuits.*

ter, and the amplified output signal is taken between collector and emitter. This configuration is analogous to the grounded-cathode tube circuit. Fig. 212-c shows the *grounded-collector* circuit, also called *common-collector*. The input signal is applied between base and collector, and the amplified output signal is taken between emitter and collector. This configuration is analogous to the cathode-follower tube circuit. Each of the three basic transistor circuits partakes of some of the characteristics of the equivalent tube circuit. Dc biasing of the surface-barrier transistor is the same as specified for p-n-p junction transistors. The emitter is positive and the collector is negative with respect to the base.

Transistor parameters

The basic parameters of a transistor are emitter voltage (v_e), emitter current (i_e), collector voltage (v_c), collector current (i_c), and base current (i_b). It is customary to represent all transistor parameters with lower-case letters, and to use capitals for external-circuit parameters. Thus in Fig. 213, v_e is the transistor emitter-to-base voltage, while V_e is the emitter-voltage supply.

Fig. 213 shows a grounded-base transistor connected directly to emitter and collector dc bias supplies, and with the basic currents and voltages indicated. For simplicity, no external-circuit elements other than the two batteries are shown. Emitter and collector

resistive parameters may be determined from the basic voltages and currents. For example, the static resistance (r_c) of the col-

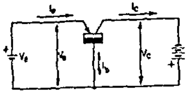


Fig 213. *Basic transistor parameters.*

lector diode $= v_c/i_c$. Emitter and collector power dissipation (P) likewise may be determined ($P = vi$).

Resistive components

With respect to internal resistance components "seen" from its three terminals, the transistor may (at dc and low audio frequencies) be represented by an equivalent three-terminal network such as is shown in Fig. 214. Actually, the network shown in this illustration is simplified and will be elaborated upon in Chapter 3. Resistances r_e and r_c are emitter and collector resistances, respectively. Resistance r_b, the base resistance, depends upon the resistivity of the germanium in the base region of the transistor and upon the resistance of the base connection, although other

Fig 214. *Simplified three-terminal resistance network of a transistor.*

factors also might govern its value to some extent. The base resistance of the high-frequency junction transistor (Fig. 208) is lowered by its peculiar shape, and is very low in the diffused-base transistor (Fig. 209).

The effect of these resistive components upon input and output resistances of the transistor in various circuit configurations may be observed from Table 2–1.

Table 2–1. Transistor Circuit Input & Output Resistance

Connection	Input Resistance	Output Resistance
Grounded-base	$r_e + r_b$	$r_c + r_b$
Grounded-emitter	$r_b + r_e$	$r_c + r_e$
Grounded-collector	$r_b + r_c$	$r_e + r_c$

Characteristic curves

Because the transistor is a current-operated device, current is taken as the independent variable in the measurement and speci-

fication of transistor characteristics. Thus, emitter or collector current is varied in test operations and the resulting emitter or collector voltage is observed. In tube technique, on the other hand, electrode voltages are varied and the corresponding currents noted.

Transistor current vs. voltage characteristics are specified with respect to a *constant-current* parameter. This is in contrast to the vacuum tube, a voltage-operated device, whose voltage vs. current characteristics are specified with respect to a constant-voltage parameter. For example: Each of a family of plate voltage vs. plate current tube curves is plotted for a constant value of grid voltage, while each of a similar set of transistor collector current vs. col-

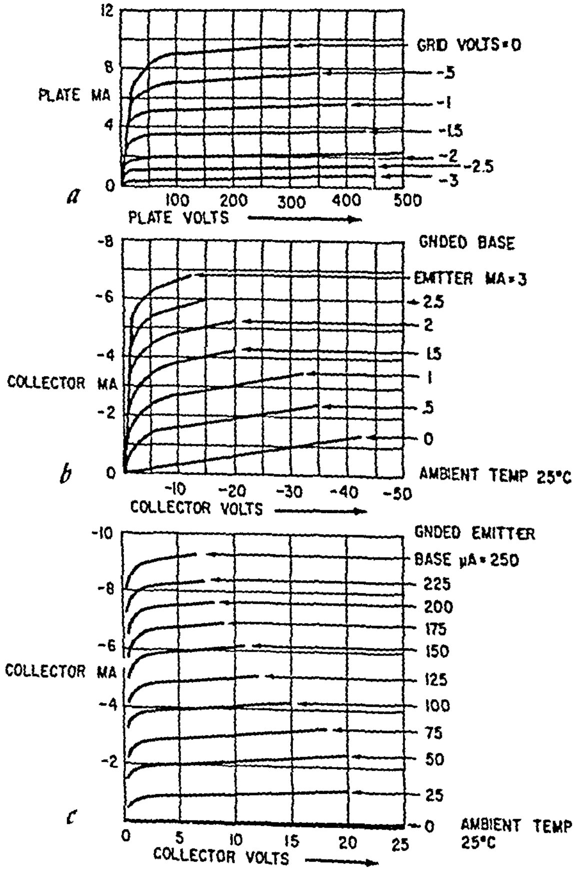


Fig. 215. *Transistor collector current vs. collector voltage curves compared with vacuum-tube characteristics.*

lector voltage curves is plotted for a constant value of emitter current (for the grounded-base connection) or base current (for the grounded-emitter connection).

Fig. 215 illustrates this point. Two sets of transistor collector current vs. collector voltage curves are shown in comparison with a set of similar tube plate voltages vs. plate current curves. Observe in Fig. 215 that the transistor collector characteristic curves (Fig. 215-b and Fig. 215-c) have the general shape of pentode tube plate curves. (Fig. 215-a). The tube curves depict variation of plate current (i_p) with plate voltage (e_p) for selected constant values of grid voltage (e_g). The transistor curves for the grounded-base circuit (see Fig. 215-b) show variation of collector voltage (v_c) with collector current (i_c) for selected constant values of emitter current (i_e). The transistor curves for the grounded-emitter circuit (Fig. 215-c) show variation of collector voltage (v_c) with collector current (i_c) for selected constant values of base current (i_b). In each of the families of transistor curves in Fig. 215, we have taken the liberty of rotating the graphs simply to give easy comparison with the accompanying family of tube curves. Properly, current values should be plotted along the horizontal axis of each transistor curve, and voltage values along the vertical axis.

Other characteristics which may be plotted for the grounded-base connection are: (1) emitter current vs. emitter voltage for constant values of collector current, (2) collector current vs. emitter voltage for constant values of emitter current, and (3) emitter current vs. collector voltage for constant values of collector current. Characteristic (2) is known as the *feedback characteristic,* since it shows the influence of output (collector) current upon input (emitter) voltage. Characteristic (3) is known as the *transfer characteristic,* since it shows the influence of input (emitter) current upon output (collector) voltage

Alpha

Current amplification factor, (emitter-to-collector) evidenced by a change in collector current for a change in emitter current, correctly refers to the current amplification between emitter input and collector output, and is a property of the grounded-base circuit. This amplification is designated by α, the Greek letter alpha.

Beta

Current amplification (base-to-collector), beta, is a second type of current amplification obtained in the grounded-emitter circuit

t is varied in test operations and the resulting emitter or or voltage is observed. In tube technique, on the other electrode voltages are varied and the corresponding currents

isistor current vs. voltage characteristics are specified with to a *constant-current* parameter. This is in contrast to the n tube, a voltage-operated device, whose voltage vs. current teristics are specified with respect to a constant-voltage para- For example: Each of a family of plate voltage vs. plate t tube curves is plotted for a constant value of grid voltage, each of a similar set of transistor collector current vs. col-

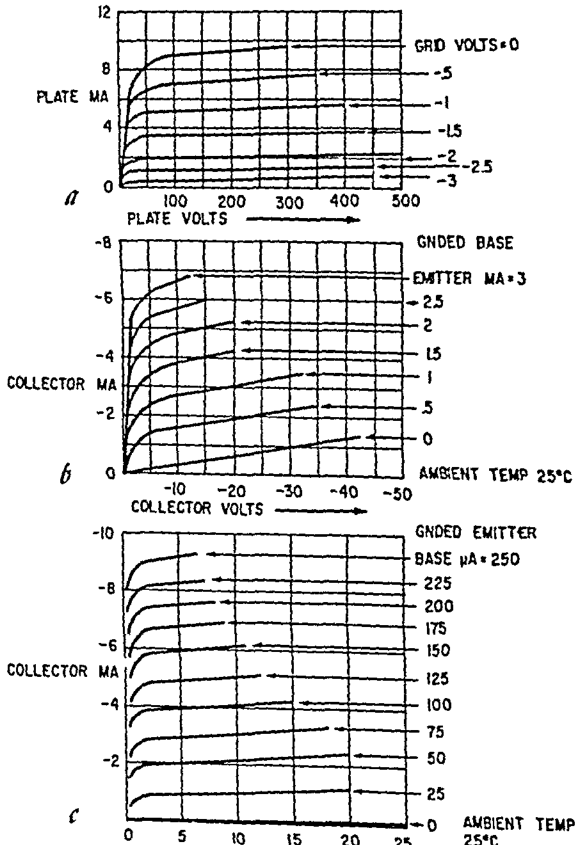


Fig. 215. *Transistor collector current vs. collector voltage curves compared with vacuum-tube characteristics.*

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Current amplification factor, (emitter-to-collector) evidenced by a change in collector current for a change in emitter current, correctly refers to the current amplification between emitter input and collector output, and is a property of the grounded-base circuit. This amplification is designated by α, the Greek letter alpha

Beta

Current amplification (base-to-collector), beta, is a second type of current amplification obtained in the grounded-emitter circuit

where the base is the input electrode. This amplification is designated by β, the Greek letter *beta*.

Beta is especially interesting in the grounded-emitter circuit employing the junction transistor, since it reaches values of 30 to 40 in many units and more than 100 in others, while alpha for the same units is less than 1. The reason for this action is that the base current is very small, compared to the collector current (i_b being the difference between collector and emitter currents). Base-to-collector amplification, beta, may be discerned readily in the grounded-emitter characteristics in Fig. 215-c. Note, for example, that at a collector voltage of 10, a base-current change of 25 microamperes (from the 25-to the 50-μa curve) results in a collector-current change of 1 milliampere (from the 1- to approximately the 2-ma ordinate). This corresponds to a beta of 40. Beta is di_c/di_b and in terms of alpha is equal approximately to $a/(1-a)$. Hence, it is desirable for high amplification in junction-transistor grounded-emitter applications that alpha be as near unity in value as possible.

Negative resistance in point-contact transistors

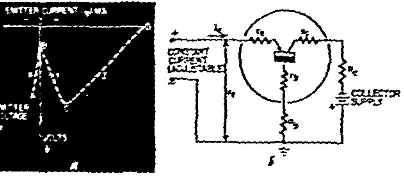
In the point-contact transistor (but not in junction types), when alpha is larger than 1 and the base resistance, r_b, is fairly high or is supplemented with an external base resistor, R_b, the transistor will exhibit negative resistance characteristics at certain levels of current and voltage. This results from positive feedback and the effect of alpha upon the voltage drop across the base resistance.

Fig. 216 shows emitter current vs. voltage relationships under these conditions. Assume the collector voltage to be held at some constant value. As the emitter current (i_e) is increased positively, the emitter-to-ground voltage (v_e) first will increase *negatively* from B to C and then will decrease from C to D. Further positive increase of i_e will cause v_e to cross the zero axis and increase beyond point D in the positive direction. If i_e is increased in the negative direction, v_e will increase negatively from B to A. The region from B to C displays negative resistance, while B–A and C–D represent positive resistance regions. The B–A region is termed cutoff, B–C negative resistance, and C–D saturation.

From the plot in Fig. 216, it is seen that while the emitter current determines the emitter voltage at all points, the emitter voltage does not determine values of emitter current in the same manner. For example; at the marked value of emitter voltage on the graph, emitter current might have either of the values x, y, or z. If the emitter current is *resting* at a lower value, the application of

a large transient will flip it to the higher value. Since a higher collector current flows in response to the higher emitter current, this unstable condition of switching to the higher current in response to large signal inputs can cause damage to the transistor unless design precautions are taken to limit the currents.

The transistor normally will be unstable in the region of B–C where its emitter-to-base input resistance is negative. However, if the emitter is biased so as to set at the stable point x along the cutoff region B–A, the application of a sufficiently large positive pulse will flip conduction to the other stable point z along the saturation region C–D. Steady emitter current, corresponding to point z, then will continue to flow even after the pulse is removed. If a strong negative pulse subsequently is applied to the emitter, the current will be flipped back to point x. This ability of the point-contact transistor to exhibit two stable states adapts it to use in flip-flop, switching, and counting circuits. The negative resistance characteristic also expedites its operation as an oscillator.



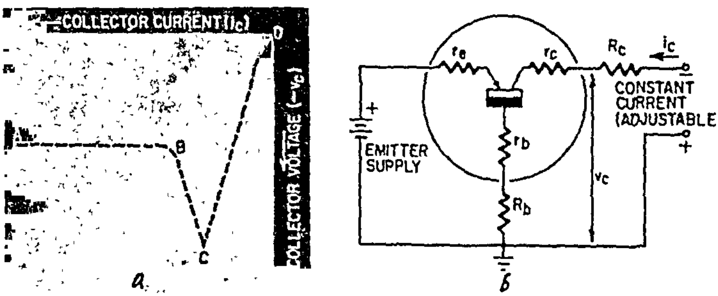
Figs. 216–a, –b. *Point-contact transistor emitter current vs. voltage characteristic.*

The mechanism of negative resistance in transistors is an extensive subject which has received considerable mathematical treatment for detailed analysis. A physical picture of how it is accomplished can be given in comparatively simple terms, however. Both emitter and collector currents flow through the base resistance and have opposite polarities. The net base current is their difference, $i_e - i_c$. In the point-contact transistor with an alpha higher than 1, the negative collector current is, by amplification, higher than the positive emitter current. The net base current thus becomes negative as long as alpha continues higher than 1, and the voltage drop across the base resistance is $v_b = r_b (-i_c - i_e)$ and is negative. Thus, the base voltage, which appears

between base and emitter, can be negative although emitter current is positive. As alpha decreases after the emitter current has reached a certain value, the negative emitter voltage reaches a turning point (C in Fig. 216) after which the positive emitter current then begins to determine the polarity of the net base current and consequently the polarity of v_b and v_e (C to D in Fig. 216).

The foregoing explanation shows that alpha must exceed unity in order for negative resistance to be evidenced. For this reason, it is easy to see why the property is not encountered in junction transistors (alpha is always less than 1).

The collector circuit also exhibits negative resistance under suitable conditions of currents and voltages, as shown by the curve in Fig. 217. Here, the emitter voltage is held at a constant value while the collector current is varied. The collector voltage first increases negatively from D to C, then decreases from C to B, and



Figs. 217-a, -b. *Negative resistance curve of collector circuit.*

finally levels off from B to A. Negative resistance is shown by the region C–B. Region D–C is cutoff, and B–A saturation.

Negative resistance is of concern both as a help and as a hindrance in the application of point-contact transistors: (1) It can be utilized in making flip-flop, switching, counting, and oscillator circuits (2) uncontrolled, it gives rise to instability in amplifier circuits; and (3) it causes negative input and output impedances. Transistors intended for stable amplifier operation are selected to have a low base resistance in the interest of minimizing tendencies toward negative resistance effects. It is obvious, however, that even a transistor so selected becomes unstable when operated in a circuit with a large-size external base resistor. *Switching type* point-contact transistors are units selected with high base-resistance values and high alphas, and are intended primarily for use in flip-flop, switching, and counting circuits.

Instability in point-contact transistors is particularly evident when input and output sources and terminations are low-impedance. The units are said to be *short-circuit unstable.* Thus, an amplifier oscillates under such conditions or when connected to low-impedance constant-voltage dc power supplies. Junction transistors are not short-circuit unstable, since they do not exhibit negative-resistance characteristics.

Collector capacitance

A space-charge region is present in the base-to-collector junction. In simple, physical terms, the presence of this region is due to the fact that all mobile carriers are removed from the zone by the collector voltage and only fixed charges due to the internal constitution of the semiconductor atoms are left. The space-charge region possesses a dielectric constant and simulates a capacitor.

This collector capacitance (C_c) is high when the space-charge region is thin, and vice versa. It is lowest when the collector-to-base reverse voltage is high. A transistor operates best at high frequencies when its C_c is low. In junction transistors designed for low-power audio-frequency applications, collector capacitance may be as [illegible]

Cutoff current

Under the heading of *Transistor action* in the first part of this chapter, it was explained that a small reverse current (i_{co}) flows when a dc voltage is applied between the collector and base and the emitter-to-base (input) circuit is open. This *cutoff* or *leakage* is a function of the collector voltage. In low-power germanium junction transistors, i_{co} varies from 1 to 10 μa, depending upon type. In silicon junction transistors, i_{co} is specified as low as .005 μa. In power transistors, i_{co} is of the order of 1 ma.

Cutoff current is temperature-sensitive. It doubles approximately for each 10° Centigrade rise in junction temperature. This is a matter of serious concern in transistor applications; since, if steps are not taken to stabilize circuit operation, i_{co} drift will change operating points and may cause collector current "run away" and destroy the transistor.

Power output

The ac output power of a transistor amplifier or oscillator, as in corresponding vacuum-type circuits, bears a percentage relationship to the dc power input supplied to the output electrode. Thus,

in a grounded-base transistor amplifier, the ac power output is a fraction of the dc collector power input.

Power output of low-power junction transistors in class-A amplifiers ranges from 1 to 25 mw, depending upon type. The same transistors will deliver up to 1/4 watt when operated as matched pairs in class-B amplifiers. Present commercial junction type *power* transistors (n-p-n and p-n-p) deliver up to 20 watts in class-A and up to 80 watts in class-B amplifiers, depending upon type. A single power transistor has been employed to switch 1,000 watts. While the cutoff frequency of most present power transistors is given as 10 to 500 kc, a laboratory-produced silicon power transistor has delivered 5 watts at 10 mc.

From a standpoint of the ratio of ac output power to dc input power (output efficiency), the junction transistor is a better performer than the vacuum tube, the transistor efficiency being up to 49% for class-A operation and as high as 75% for class B. All transistors may be considered more efficient than tubes as power converters if overall power supply requirements are considered, since the transistor requires no filament power.

In grounded-base and grounded-emitter circuits, power output is limited by allowable collector power dissipation. In the grounded-collector circuit, where the emitter is the output electrode, the limiting factor is the maximum allowable emitter power dissipation.

There is a maximum amount of power which can be dissipated by either the emitter or collector without damage to the transistor or a change of its characteristics through heating. Emitter and collector ratings vary with types and manufacture. The permissible

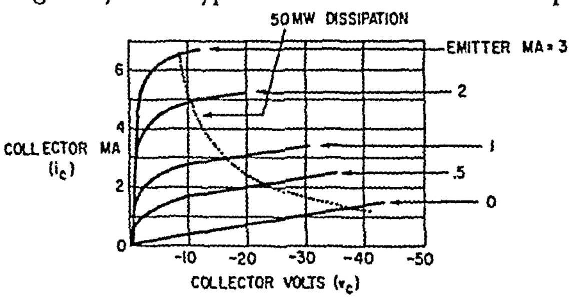


Fig. 218. *Collector current vs. collector voltage curves.*

emitter dissipation always is lower than the collector dissipation. Typical maximum values are 50 to 150 mw in point-contact types, 10 to 200 mw in low-power junction types, and 3 to 60 watts in

From a practical standpoint, it is important to note that maximum rated current and voltage do not always coincide with maximum power dissipation in all circuits. A great deal depends upon the effect of a third parameter, such as emitter or base current, when collector dissipation is considered. Thus, under circuit operating conditions, maximum rated dissipation might be exceeded long before reaching maximum rated current or voltage.

Fig. 218 gives a family of collector current vs. collector voltage curves for five constant values of emitter current. A line has been dotted in to show 50-mw collector dissipation. From the curves, note that the 50-mw dissipation is not exceeded at a collector voltage of −30 when the emitter current is zero, but is exceeded at this voltage easily when the emitter current is 0.25 or more ma.

Emitter dissipation is not always listed in tables of transistor operating data. Instead, maximum emitter current is shown. The operator must exercise care to keep within this rating.

Power gain

The transistor amplifier resembles the class-B vacuum-tube amplifier in the respect that each requires a finite amount of signal input power. This is grid-signal power in the tube circuit, and is emitter input or base input in the transistor amplifier.

The overall power gain (ratio of output-signal power to input-signal power) provided by the transistor depends upon the type of circuit connection employed. Typical values are 18 to 25 db for the high-grade point-contact type in a grounded-base circuit, 30 to 40 db. for high-alpha junction types in a grounded-emitter circuit, and 12 to 18 db for high-alpha junction types in a grounded-collector circuit. Transistor power gain varies approximately as the ratio of output to input impedance, and also as the square of alpha.

Noise

Transistor noise level is higher than that of a vacuum tube and increases in importance as the number of cascaded amplifier stages is increased. The noise figure in transistor operating data is specified as so many decibels above thermal noise at 1,000 cycles. Typical noise-figure values for junction transistors vary between 6 and 30 db. The figure usually is based upon grounded-emitter operation at a specified dc collector voltage and 1-cycle bandwidth at 1,000 cycles. For a given transistor, noise decreases with operating frequency and is higher when it originates at the collector than at the emitter.

frequency response

A combination of frequency-dependent factors, such as current amplification, voltage amplification, internal capacitances, transit time, point-contact spacing (close spacing for high-frequency response), width of base region, spreading of carriers, and equivalent network impedances, act to limit the maximum frequency at which satisfactory operation can be obtained.

In the point-contact transistor, transit time, upon which frequency response depends, varies inversely as the square of the contact spacing. Point-contact transistors made with p-type germanium have somewhat better frequency response than those made with n-type. The reason for this is the higher mobility of electrons which are the carriers in p-type germanium. Holes, the carriers in n-type germanium, travel much more slowly.

In the junction type, the high-frequency operating limit can be extended by making the central base-region layer extremely thin to reduce transit-time effects. Special constructions, such as those shown in Figs. 207, 208, and 209, also increase the operating frequency range. Contemporary diffused-base transistors are rated for operation beyond 100 mc, and surface-barrier transistors to 60 mc. Power type triode transistors have alpha cutoff frequencies up to 0.5 mc. A contemporary high-power tetrode transistor has a beta cutoff-frequency rating of 12 kc. Experimental diffused-base transistors in the laboratory have been reported to operate as high as 600 mc.

The conventional manner of rating transistor frequency response is to specify a *cutoff frequency*, the frequency at which alpha drops to a point 3 db below its low-frequency value. Some manufacturers specify beta rather than alpha, and still others specify the frequency at which amplification through the transistor drops to unity.

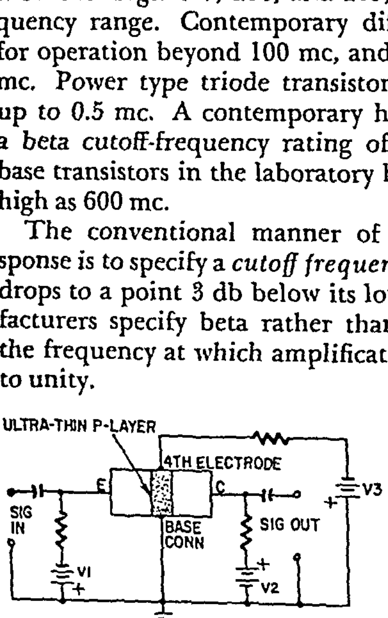


Fig. 219. *High-frequency junction tetrode transistor.*

The n-p-n junction *tetrode* transistor of R. L. Wallace (see Fig. 219) overcomes the effect of high base resistance by means of a negative dc bias applied to the central p-layer at a point opposite

the base contact. This bias repels the electrons coming from the emitter, forcing them to diffuse through the p-layer near the base connection, instead of all through the layer, thereby reducing the base resistance as much as 10 times.

Contemporary low-power tetrode transistors are rated for 120-mc operation.

Temperature effects

The temperature dependence of germanium was stressed in Chapter 1. In the transistor, elevated temperatures act to reduce electrode resistances, decrease amplification and increase noise. Changes due to temperature appear somewhat more severe in junction transistors than in point-contact units. Operating characteristics usually are specified for 25°C. Typical maximum ambient operating temperature is 70°C (158°F) for germanium and 175°C (347°F) for silicon transistors.

The drastic effect of temperature in increasing the collector leakage current, i_{co}, was mentioned earlier under *Cutoff current*. This current is higher in the grounded-emitter circuit than in the grounded-base because of the high base-to-collector amplification factor (beta) of the former which tends to magnify the current. I_{co} drift with temperature therefore is much more troublesome in the grounded-emitter circuit.

Life

Because of the comparative infancy of the transistor, extensive life data comparable to that available for vacuum tubes and semiconductor diodes have not yet been accumulated. However, tests indicate that life ratings in multiples of 10,000 hours can be expected. Authorities predict a figure of 70,000 hours, which means that a transistor under favorable conditions might operate continuously for 8 years.

Comparison of dc requirements

In commercial low-power transistors (depending upon model and manufacture), maximum dc collector current ratings vary from 2 to 20 ma for the point-contact type and from 5 to 500 ma for the junction types. Maximum dc collector voltages vary from 10 to 100 for the point-contact type and from 20 to 105 for junction types.

In commercial power junction transistors (depending upon model and manufacture), maximum dc collector current ratings vary from 500 ma to 13 amperes, and maximum dc collector voltages from 6 to 80.

A useful property of the low-power junction transistor is its ability to operate at very low levels of voltage and current. Practical oscillators have been demonstrated to operate on direct current generated by a photocell, a thermocouple, or a pair of coins separated by moistened paper. This property has made possible the design of subminiature amplifiers, radio receivers, and hearing aids that may be operated from inexpensive 1.5-or 3-volt batteries.

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equivalent circuits

BECAUSE of the vast amount of experience gained with vacuum tubes in electronic circuits and the relatively limited amount of similar experience with transistors, the general tendency is to think of transistors as simply replacing tubes in common circuits. This is not a correct attitude nor a practical one, since there is a basic difference in nature between the control of current carriers in a vacuum and in a solid.

Unlike the vacuum tube, the transistor *does not* provide isolation between its input and output circuits. Because of this, output conditions in a transistor affect input parameters and vice versa. The transistor resembles a set of connected resistances. It is, in fact, an *active* resistance network. The successful application of this device to electronic circuits must be based upon its nature as a network.

At dc and low audio frequencies, the transistor resembles, and may be described in terms of a three-terminal resistance network. This configuration was shown in simplest terms in Chapter 2. We are in a position now to consider the transistor network in more detail. Transistor circuits may be described also in terms of four-terminal networks, but we shall restrict most of our discussion to the three-terminal variety which serves the purposes of this book.

Network parameters

For the benefit of readers who have no background of network theory, the following definitions of network terms are made with respect to transistor parameters with which they are associated.

For reference, Fig. 301 shows the equivalent three-terminal resistance network of a transistor operating as a grounded-base amplifier. The purely resistive components are allowable for small-signal conditions at dc and low audio frequencies. These become impedances, with reactive components, at high frequencies.

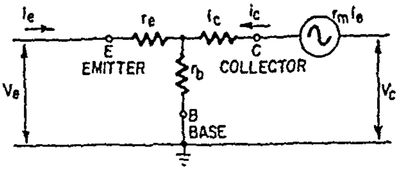


Fig. 301. Equivalent circuit of a transistor grounded-base amplifier at dc and low audio frequencies.

Only the transistor impedances are shown. External circuit components, such as load and generator, have been omitted from Fig. 301 for simplicity, but appear in succeeding diagrams.

For the grounded base

(3–1) R_{11} = input resistance = slope of the curve v_e vs. i_e when i_c is constant $= \left.\dfrac{dv_e}{di_c}\right| i_c$

(3–2) R_{12} = reverse transfer resistance = slope of the curve v_e vs. i_c when i_e is constant $= \left.\dfrac{dv_e}{di_c}\right| i_e$

(3–3) R_{21} = forward transfer resistance = slope of the curve v_c vs. i_e when i_c is constant $= \left.\dfrac{di_e}{dv_c}\right| i_c$

(3–4) R_{22} = output resistance = slope of the curve v_c vs. i_e when i_e is constant $= \left.\dfrac{dv_c}{di_c}\right| i_e$

(3–5) alpha $= \alpha =$ current amplification factor = slope of the curve i_c vs. i_e when v_c is constant =

$$\left.\frac{di_c}{di_e}\right| v_c = R_{21}/R_{22}$$

(3–6) r_m = active mutual characteristic of network $= R_{21} - R_{22} = \alpha r_c$

The output generator in the collector lead in Fig. 301 has a value $r_m i_e$ which corresponds to the generator μE_g in the equivalent circuit of a vacuum-tube voltage amplifier. The active properties of the transistor network are expressed by this generator.

These equations are for the grounded-base amplifier. Similar equations are used to describe the network characteristics of other circuit configurations; e.g., the grounded-emitter and grounded-

collector circuits, as will be shown later. However, certain terms have the same meaning in each circuit. For example; R_{11} is always the input resistance, R_{22} the output resistance, R_{12} reverse transfer resistance, R_{21} forward transfer resistance, and r_m the active mutual resistance.

Grounded-base amplifier

Fig. 302 shows the grounded-base amplifier circuit with its equivalent three-terminal network. Bias supplies have been omit-

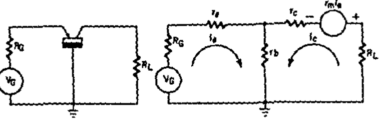


Fig. 302. Grounded-base amplifier and equivalent three-terminal network.

ted. V_G is the source generator with internal resistance R_G. Component R_L is the load resistance.

(3–7) $r_m = \alpha (r_e + r_b) - r_b$. From Equation (3–6), r_m also $= \alpha r_c$

(3–8) $R_{11} = r_e + r_b - \dfrac{r_b (r_b + r_m)}{R_L + r_c + r_b}$

(3–9) $R_{12} = r_b$

(3–10) $R_{21} = r_b + r_m$

(3–11) $R_{22} = r_c + r_b - \dfrac{r_b (r_b + r_m)}{R_G + r_e + r_b}$

Grounded-emitter amplifier

Fig. 303 shows the grounded-emitter amplifier circuit with its

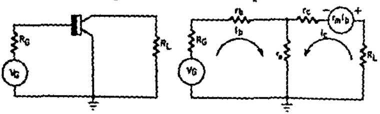


Fig 303. Grounded-emitter amplifier and its equivalent network.

equivalent network. Note that the output generator here is equal to $r_m i_b$, since the input current is the base current, i_b.

(3–12) $R_{11} = r_e + r_b + \dfrac{r_e (r_m - r_e)}{R_L + r_e + r_c - r_m}$

(3–13) $R_{12} = r_e$

$$(3\text{–}15) \quad R_{22} = r_c + r_e - r_m + \frac{\ldots}{R_G + r_b + r_e}$$

Grounded-collector amplifier

Fig. 304 shows the grounded-collector amplifier circuit with its

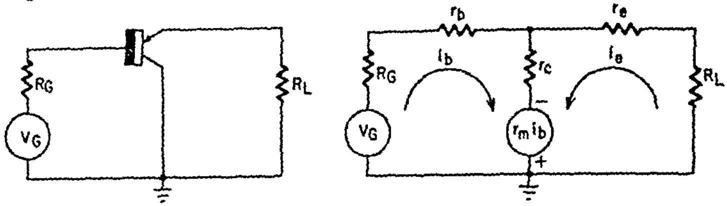


Fig. 304. Grounded-collector amplifier and its equivalent network.

equivalent network. Here, the emitter is the output electrode and the generator $r_m i_b$ is in the collector lead.

$$(3\text{–}16) \quad R_{11} = r_b + r_c + \frac{r_c\,(r_m - r_c)}{R_L + r_e + r_c - r_m}$$

$$(3\text{–}17) \quad R_{12} = r_c$$

$$(3\text{–}18) \quad R_{21} = r_c\,(1 - \alpha)$$

$$(3\text{–}19) \quad R_{22} = r_e + r_c - r_m + \frac{r_e\,(r_m - r_c)}{R_G + r_b + r_c}$$

Power amplification

The operating power gain (G) of transistor amplifier stages is closely related to values of the network components.

Grounded base

$$(3\text{–}20) \quad G = 4R_GR_L \left[\frac{-(r_b + r_m)}{(R_G + r_e + r_b)\,(R_L + r_c + r_b) - r_b\,(r_b + r_m)}\right]^2$$

For stability, the denominator of the fraction must be greater than zero.

Grounded emitter

(3–21) $G = 4R_GR_L$ (multiplied by the quantity shown below)

$$\left[\frac{r_m - r_e}{(R_G + r_b + r_e)\,(R_L + r_e + r_c - r_m) + r_e\,(r_m - r_e)}\right]^2$$

For stability, the denominator of the fraction must be greater than zero.

Grounded collector

(3–22) $G = 4R_GR_L$ (multiplied by the quantity shown below)

$$\left[\frac{-r_c}{(R_G + r_b + r_e)\,(R_L + r_e + r_c - r_m) + r_c\,(r_m - r_c)}\right]^2$$

For stability, the denominator of the fraction must be greater than zero.

Reverse power amplification

The grounded-collector *point-contact* amplifier has the attractive property of bilateral amplification when alpha is greater than 1. When $\alpha = 2$, the power gain from output to input is the same as from input to output. Reverse or "backward" power gain G_R, may be expressed in terms of network parameters:

$$(3\text{-}23) \quad G_R = \alpha R_g R_L \left[\frac{-r_c + r_m}{d}\right]^2$$

d is the denominator from Equation (3–22).

Reverse power gain may be determined also by multiplying the forward power gain by $(1-\alpha)^2$.

Electrode external resistances

The terms r_b, r_c, and r_e appearing in the equivalent circuits represent the *total* resistance associated with the base, collector, and emitter electrodes, respectively. This includes any external resistance connected in the leads, as well as the internal resistance determined by v/i. External series resistances occasionally are connected in the leads for purposes of stabilization or feedback. Not included, however, is the output load resistance, which is designated R_L in the network equations and is treated separately from r_b, r_c, and r_e.

Equivalent noise circuit

Fig. 305 is the equivalent circuit of a transistor with emitter and collector noise sources represented by the noise generators n_e

Fig. 305. Equivalent circuit of a transistor, showing various noise sources.

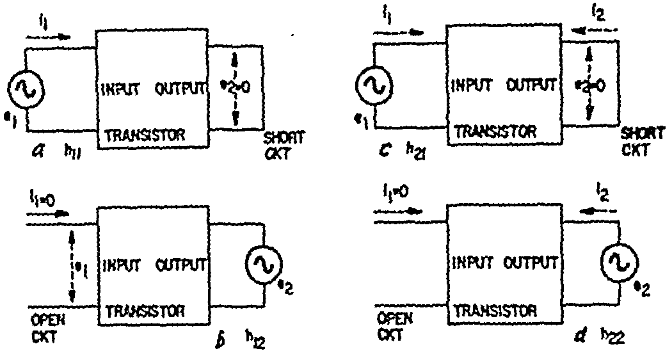
and n_c, respectively. The noise generators retain their positions also in the grounded-emitter and grounded-collector amplifier circuits, but the generator $r_m i_e$ changes position as indicated in Figs. 303 and 304.

Hybrid parameters

Instead of the r parameters (R_{11}, R_{12}, R_{21}, and R_{22}) previously described, some transistor manufacturers use hybrid (h) para-

eters in their specifications. In Fig. 306 the transistor is viewed as a four-terminal network.

In Fig. 306-a, the output is short-circuited and the output voltage, e_2, therefore, is zero: $h_{11} = de_1/di_1$. In Fig. 306-b, the input is open-circuited: $h_{12} = de_1/de_2$. In Fig. 306-c, the output is short-circuited and the output voltage e_2, therefore, is zero: $h_{21} = di_2/di_1$. In Fig. 306-d, the input is open-circuited: $h_{22} = di_2/de_2$.



Figs. 306-a to d. *Setups for hybrid parameters.*

Parameter h_{11} is called *input resistance with output short-circuited*, h_{21} *short-circuit current gain*, h_{12} *voltage feedback ratio with input open-circuited, and* h_{22} *output admittance with input open-circuited*. The h parameters bear the following relationships to the grounded-base internal resistances and current amplification factor:

$$(3\text{–}24) \quad h_{11} = r_e + r_b(1 - \alpha)$$
$$(3\text{–}25) \quad h_{12} = r_b / (r_b + r_c)$$
$$(3\text{–}26) \quad -h_{21} = \alpha$$
$$(3\text{–}27) \quad h_{22} = 1 / (r_b + r_c)$$

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transistor amplifiers

THE basic building blocks of transistor amplifiers are t single-stage grounded-base, grounded-emitter, and grounde collector circuits. The chart in Table 4–1 compares the practi operating features of each type of circuit separately for point-co tact and low-power junction transistors. This chart has been p pared from condensed characteristics of transistors available at t time of compilation. The listed values represent the spread show by the published characteristics. These values are subject to co siderable future variation from those listed, since transis manufacture is by no means fixed at this time.

Which circuit and type of transistor are selected for a given a plication will depend largely upon desired input and output i pedances, power gain, power output, and frequency respon These characteristics are located quickly in Table 4–1. Input a output parameters are shown as resistances and apply only to lo frequency operating conditions. Power output listings are t maximum values specified by transistor manufacturers

The chart data are from *typical* operating conditions and a subject to variation under different conditions of application. T equivalent circuits in Chapter 3 show how input and output sistance and power gain are mutually dependent.

From Table 4–1 several important aspects of transistor amp fiers are apparent: (1) Only one circuit, the grounded-emitt reverses the phase of the transmitted signal (2) The ground emitter circuit also provides the highest power gain, although n the greatest power output. (3) The lowest input-resistance val

occur in grounded-base circuits. (4) The highest input resistance is provided by the grounded-collector circuit, although this resistance is dependent *directly* upon the value of output (load) resistance used.

From a practical viewpoint, junction transistors of both types can be used successfully in any of the three circuits, while point-contact units give good performance only in the grounded-base circuit. While it is possible to operate the point-contact type in the other two circuits, its behavior would be somewhat inferior to that of the junction type because of instability of the contact type in those circuits and because of the limited permissible dissipation of the emitter which becomes the output electrode.

Table 4—1. Comparison of Basic Transistor Amplifiers

Circuit	Transistor Type	Input Resistance (ohms)	Output Resistance (ohms)	Power Gain (db)	Power Output (mw)	Phase Reversal
Grounded-Base	Point-Contact	100–400	6,000–40,000	19–21	30–60	No
	Junction	10–300	100,000–500,000	20–29	2–15	No
Grounded-Emitter	Junction	325–1,000	5,000–40,000	30–40	2.8–20	Yes
Grounded-Collector	Junction	150,000–800,000	200–20,000	12	—	No

Figs. 401, 402, and 403 show typical single-stage amplifiers using grounded-base, grounded-emitter, and grounded-collector circuits. In each figure, *a* designates the R-C-coupled version of the circuit, while *b* shows the transformer-coupled version. The battery polarities are correct for n-type point-contact and p-n-p junction transistors. All battery polarities must be reversed for n-p-n junction types and p-type point-contact transistors. Otherwise, the circuits are the same.

Single vs. multiple bias supply

In each of the circuits in Figs. 401, 402, and 403, separate bias supplies are shown for emitter and collector. While batteries are shown for simplicity, the supplies might also be of the ac rectifier type.

The dual supply scheme is not the only practical method of operation. A single battery is sufficient in many applications and often is attractive from a standpoint of economy, simplicity, and

ance, but large input signals can be handled before peak clipping sets in. Fig. 404-c is a grounded-base circuit in which emitter bias is developed across the external base resistor, R_b. This action is somewhat analogous to self-bias with a cathode resistor in a tube circuit. The use of a base resistor in this manner is not recommended for point-contact transistors, since it encourages instability and oscillation as the result of positive feedback. (See the discussion of *negative resistance* in Chapter 2.) Figs. 404-d, -e show grounded-collector circuits corresponding to the two types of single-battery, grounded-emitter circuit in Figs. 404-a and -b. The remarks already made regarding input impedance and signal level for the floating-base and biased-base grounded-emitter stages also apply to floating-base and biased-base grounded-collectors.

Constant-current requirement

The voltage-operated nature of the vacuum tube conventionalized the low-impedance, constant-voltage power supply. The transistor favors constant-current dc power supplies. We have already seen that the point-contact type, at least, is short-circuit unstable. Its power supply therefore must be a *high-impedance* type.

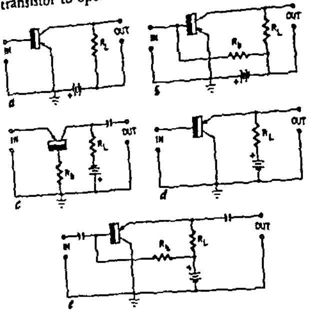
The simplest constant-current power supply is a high voltage connected in series with a high resistance. The series resistance should be higher than the self-resistance of the transistor electrode supplied. In the amplifier circuits in Figs. 401, 402, and 403, the circuit resistances (R_b, R_c, and R_e) in the circuits labeled *a* supply constant current if they and the battery voltages (V_c and V_e) are high enough. In the circuits labeled *b*, suitable constant-current resistors must be connected in series with the input and output electrodes if the transistors are short-circuit unstable.

The use of constant-current bias supplies minimizes variations in circuit performance between individual transistors and, to some extent, those variations due to temperature, since the large external resistance, rather than the transistor self-resistance, determines the electrode current.

Low-power operation

To the increased operating efficiency of the junction transistor is added its attractive feature of operating satisfactorily at low voltages and currents. Practical amplifiers and oscillators using junction transistors can operate at collector voltages as low as 1.5 and 3, and at currents of less than 100 μa. Experimental oscillators have been operated at a few millivolts of collector potential and

a few microamperes of collector current. This ability of the junction transistor to operate at low power levels results in great



Figs. 404-a to e Typical single-battery amplifier stages.

economy and compactness in portable and sub-miniature electronic equipment. No equal performance is possible with present vacuum tubes.

Input–output coupling

Input and output coupling for a transistor amplifier stage may be of the capacitive, transformer, or direct type. Figs. 401, 402, and 403 give examples of capacitive and transformer input and output coupling, while Figs. 404-a, -b, -c, and -d show direct coupling. The spread of input and output impedances (resistances) typical of the three types of amplifier stages may be ascertained from Table 4–1.

Capacitive coupling may be employed successfully when the signal-source impedance is comparable to, or lower than the transistor input resistance (R_{11}) and when the impedance of the output device is comparable to, or higher than the output impedance (resistance) of the transistor (R_{22}). Thus, a low-impedance dynamic or carbon microphone might be operated directly into the R-C-coupled transistor input, while a high-impedance crystal pickup could *not*. Similarly, an ac vacuum-tube voltmeter with its high impedance characteristic could be driven directly from

the R-C-coupled output of a transistor stage, while low-impedance headphones could not.

Transformer input and output are required whenever the impedances of the signal source and load device differ significantly from the transistor input and output resistances, respectively, and impedance matching thus is essential. Also, higher power gain per stage is obtained with transformer coupling.

Direct coupling is employed for dc amplification or when coupling components, such as capacitors or transformers, must be omitted from the circuit. In direct coupling also, generator and load impedances must have the proper magnitude relationships to the transistor input and output impedances. That is, the generator impedance must be comparable to, or smaller than the transistor input resistance—and the impedance of the load device must be comparable to, or higher than, the transistor output resistance.

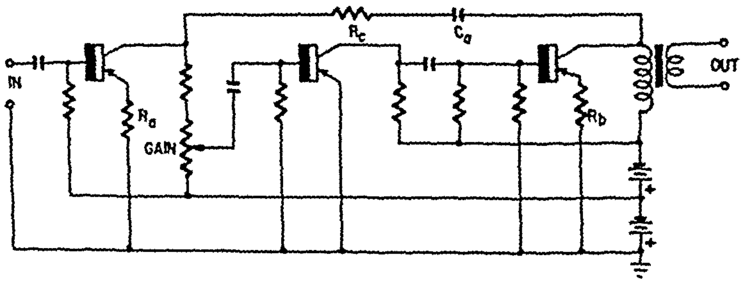


Fig. 405. *Resistance-coupled three-stage transistor amplifier.*

In some applications, it is desirable to connect the load device directly in series with the output electrode of the transistor. Examples of load devices which may be connected in this manner are dc meters, high-resistance headphones, dc relays, high-impedance speakers, and neon lamps. Such connection is permissible when the steady component of output electrode current will not interfere with normal operation of the device and when the dc resistance of the device is not so high as to reduce significantly the electrode voltage.

Cascading amplifier stages

Transistor amplifier stages can be cascaded, following, in general, the same schemes employed in multistage vacuum-tube amplifiers. Any desirable combination of grounded-base, grounded-emitter, and grounded-collector stages may be cascaded to increase

the R-C-coupled output of a transistor stage, while low-impedance headphones could not.

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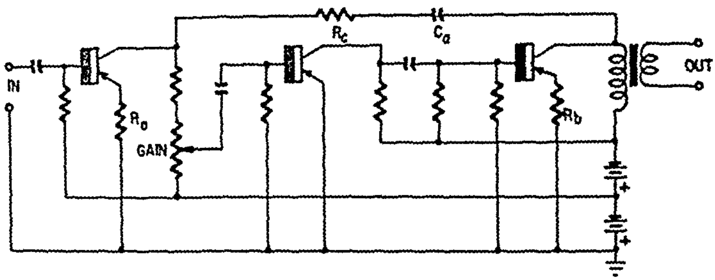

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Cascading amplifier stages

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voltage gain or power gain. Noise level is the factor which usually limits the number of stages which can be operated practically.

Cascading transistor amplifier stages is not as easy a matter, however, as building a multistage vacuum-tube amplifier because, in every case except the grounded-collector, the transistor input impedance is lower than the output impedance of a preceding transistor stage. Appreciable power is lost in interstage coupling unless impedance matching is employed. Maximum possible gain per stage therefore cannot be realized in simple R-C-coupled cascades such as shown in Fig. 405, although the sacrifice in gain

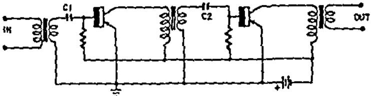


Fig. 406. Transformer-coupled two-stage amplifier.

sometimes is justified by the increased simplicity of this arrangement. A rule of the thumb which has arisen in transistor amplifier practice is that at least one more stage is required for a given overall gain with R-C-coupling than would be needed if interstage transformer coupling, such as shown in Fig. 406, were used.

From Table 4—1, it is seen that cascaded, grounded-base, junction stages could necessitate working between a 100,000-ohm output and 10-ohm input impedance. This impedance mismatch in R-C coupling would cause such a gain loss that *several* such stages might be needed to give an overall gain equal to the total gain of just two transistors. Only by using interstage transformers to supply the stepdown impedance match would cascading these grounded-base transistors be feasible.

Interstage transformers for use between any combination of grounded-base and grounded-emitter stages must provide stepdown ratios. Only when working out of any transistor stage into a grounded-collector stage is a stepup ratio required.

The grounded-collector stage is used occasionally as an impedance matcher between two other transistor stages, especially when complete use of R-C coupling is required. In this way, a fair transfer of energy may be obtained between the stages separated by the grounded-collector amplifier. As has been pointed out in earlier chapters, the grounded-collector amplifier will provide fair power gain, but its voltage amplification never can exceed unity.

he grounded-collector may be employed as the input stage to ve a transistor amplifier relatively high input impedance.

Because the input and output resistances of the transistor in grounded-base and grounded-emitter circuits are relatively low, the coupling capacitors in Figs. 405 and 406 must be high in value. For good low-frequency response, these capacitances seldom are under 1 μf each and preferably should be 10 to 20 μf. The

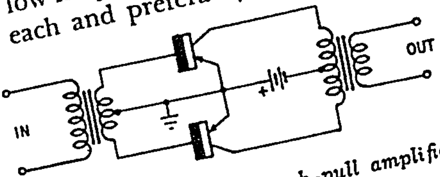


Fig. 407. *Conventional push-pull amplifier.*

capacitors (C1 and C2) are required in the transformer-coupled circuit (Fig. 406) to prevent grounding of the base bias through the transformer secondary windings.

While only two typical examples of cascaded-stage transistor amplifiers are shown here, multistage transistor amplifiers to satisfy individual requirements can be built with various combinations of R-C and transformer coupling and different combinations of grounded-base, grounded-emitter and grounded-collector stages.

Conventional push-pull operation

Transistor pairs can be operated in conventional push-pull am plifier circuits, as shown in Fig. 407. However, close matching o the transistor characteristics is necessary for efficient push-pull o eration with lowest distortion.

Although Fig. 407 shows the grounded-emitter connection each half of the push-pull circuit, grounded-base and ground collector configurations also are usable in push-pull amplifiers

For input and output impedance matching, the two halve the circuit may be treated as separate transistor amplifiers. T the upper half of the input transformer secondary matche impedance of the upper-transistor input, while the lower h the secondary matches the input resistance of the lower tran The same applies to the output transformer.

Push-pull by complementary symmetry

Fig. 408 shows an ingenious single-ended push-pull a developed by George C. Sziklai of RCA Laboratories, util complementary symmetry of the characteristics of n-p-n a

junction transistors. In achieving push-pull operation, this circuit operates from a single-ended input and requires neither a phase inverter nor input and output coupling transformers.

Operation is based upon the fact that base-current changes in the *same* direction in an n-p-n and a p-n-p transistor will cause collector-current changes in opposite directions in the two units. Thus, collector current will rise in one unit and fall in the other.

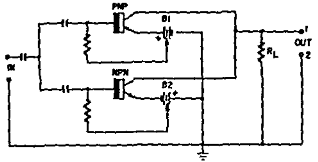


Fig 408. Push-pull amplifier using complementary symmetry principle.

When an input signal is applied to the bases of an n-p-n and a p-n-p transistor in parallel, as in Fig. 408, the transistors deliver amplified output in true push-pull fashion.

On positive half-cycles of the input signal, reduction of the emitter-collector resistance of the n-p-n unit (due to increased base current in this transistor) causes output terminal 1 to swing negative, approaching the potential of the dc supply B2. Simultaneously, the positive half-cycle is applied to the p-n-p unit, reducing the base current of the latter transistor and raising its emitter-collector resistance.

On negative half-cycles of the input signal, the base current of the p-n-p unit increases, reducing the emitter-collector resistance of this transistor and causing output terminal 1 to swing positive, approaching the potential of the dc supply B1. At the same time, the emitter-collector resistance of the n-p-n unit is increased because of reduced base current in this latter transistor.

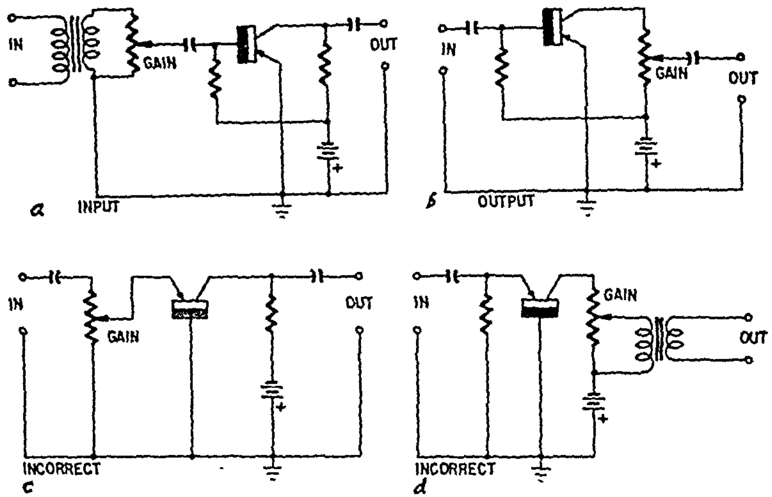
In one practical audio application, the high-impedance voice coil of a speaker has been substituted for the load resistor, R_L.

Push-pull class-B operation

The grid bias in a class-B vacuum-tube amplifier always is set for plate-current cutoff in the absence of an input signal. In a class-B transistor amplifier, emitter bias or base bias may be set to

obtain either collector-current cutoff or collector-voltage cutoff, under zero-signal conditions.

It usually is more desirable to operate at collector-current cut-off, since the other mode of operation necessitates high collector-current drain at zero signal—a condition of low resting efficiency.



Figs. 409-a to d. *Positions of the gain control in amplifier circuits.*

Each transistor delivers a single half-cycle of output voltage when driven by the relatively large input signal.

At a given power level, push-pull class-B transistor amplifiers have been found to be more efficient than similar vacuum-tube amplifiers. Output efficiencies in excess of 80% have been obtained in the laboratory, with output power from 400 milliwatts to 2 watts in the af spectrum. Contemporary power transistors, operated class-B, deliver up to 80 watts of output.

Position of the gain control

In a transistor amplifier, the gain control must be connected so that minimum disturbance is caused to any impedance match in the circuit. Some of the conventional positions of gain-control potentiometers in tube circuits therefore would not be satisfactory in transistor circuits.

Fig. 409 shows several possible positions of the gain control. The schemes in Figs. 409-a and -b are satisfactory. The input impedance of the transistor is not altered by settings of the potentiometer in Fig. 409-a nor the output impedance by the control in

Fig. 409-b. This is not true of the other two illustrations. In Fig. 409-c, the control setting inserts more or less resistance between the emitter and base, varying the input impedance of the transistor. In Fig. 409-d the potentiometer settings destroy the match between the collector and the primary of the output transformer.

Negative feedback

Feedback may be introduced into transistor amplifier circuits by methods similar to those employed in tube circuits and comparable advantages may be obtained. Negative feedback is obtainable only in those circuits in which the phase of the transmitted signal is reversed properly. This is an important point

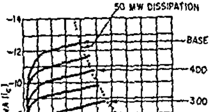


Fig. 410. Family of transistor curves corresponding to 11 base-current values.

when anticipating the application of degeneration since, as was explained earlier, not all transistors reverse signal phase

Overall feedback is secured in the three-stage transistor amplifier shown in Fig. 405 through capacitor C_a and resistor R_c. Individual-stage feedback is obtained through the emitter series resistors, R_a and R_b.

A condition to be considered when applying overall (output-to-input) feedback in transistor amplifiers is the low-impedance input circuit. This is quite different from the condition of extremely high grid-input impedance in vacuum-tube circuits. The low input impedance of the transistor causes some loss of feedback voltage by voltage-divider action or impedance mismatch. The feedback voltage must be taken from a point of sufficiently high potential to compensate for this unavoidable voltage division.

Collector output charactor

For graphic manipulation tics, as displayed by a famil rent curves, may be han vacuum tube E_p I_p curves

Fig. 110 shows a family to 11 base current values tron type 2N37 p-n-p junc 1,000-ohm load line and sponding to the 50-mw ra facturer. Class A operation graph where the curves sp Note that all values for the mw dissipation range

The slope of the [illegible] dv di [illegible] In Fig. 110 this [illegible]

Direct-coupled amplifier

[illegible]

emitter followed by a grounded base. P-n-p junctions are assumed. Battery B1 has correct polarity for supplying the two emitters, and B2 for the two collectors. This system has lower overall gain than the one shown in Fig. 411 because of the reduced gain of the grounded-base stage.

Transistors in rf amplifiers

The foregoing discussions have been mainly concerned with af amplifiers. Transistor amplifiers may be operated also as radio

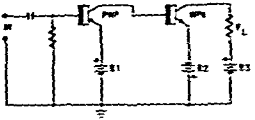


Fig. 411. Direct-coupled amplifier using p-n-p and n-p-n transistors.

frequencies, provided the transistors themselves will reach the frequencies of interest.

Many factors limit transistor frequency range. One is the diffusion time of electrons and holes in the semiconductor. Reactive

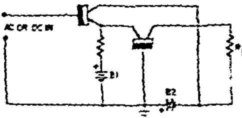


Fig. 412. Direct-coupled amplifier using identical transistor types

effects are evident at high frequencies. Collector resistance, emitter resistance, emitter and collector capacitances, α and r_m show pronounced frequency dependence beyond the audio spectrum. Circuits and operating points also influence frequency range. The upper frequency increases, for example, with collector voltage. In general, the grounded-collector circuit gives the highest frequency response. Grounded-emitter bandwidth in junction transistors is improved somewhat when the generator and load resistances are reduced, but this is at the expense of power gain.

Collector output characteristics

For graphic manipulations, the static collector dc characteristics, as displayed by a family of collector voltage vs. collector current curves, may be handled in much the same manner as vacuum-tube E_p–I_p curves.

Fig. 410 shows a family of v_c–i_c transistor curves corresponding to 11 base-current values. These curves are data for the CBS-Hytron type 2N37 p-n-p junction transistor. We have drawn in a 1,000-ohm load line and a collector dc dissipation line corresponding to the 50-mw maximum value specified by the manufacturer. Class-A operation is restricted to that portion of the graph where the curve separations are equal (linear operation). Note that all values for the 1,000-ohm load are well inside the 50-mw dissipation range.

The slope of the load line expresses its resistance. Slope $= dv_c/di_c = (v_1-v_2)/(i_1-i_2)$ where v_c is in volts and i_c in amperes. In Fig. 410, this equals $8-0/.008-0 = 1{,}000$ ohms. At any point along the dissipation line, the product $v_ci_c = 50$ mw. $(V_{cb})-V$ at the bottom of the graph, Fig. 410, simply means that the abscissas are in minus volts, and that this is particularly collector-to-base volts (V_{cb}).

Direct-coupled amplifier

Numerous methods are available for direct coupling between cascaded transistor amplifier stages. Fig. 411 shows one system, originated by G. C. Sziklai of RCA Laboratories, using a p-n-p transistor followed by an n-p-n unit.

Amplified collector output current of the p-n-p transistor flows through the base–emitter circuit of the n-p-n unit, constituting the signal-input current of the latter. This current then is amplified further by the n-p-n transistor. The first transistor being directly connected to the second one, the problem of interstage impedance matching is removed. The p-n-p unit receives its negative collector voltage from the n-p-n emitter battery, B2, through the emitter–base path of the n-p-n unit.

By making use of the complementary symmetrical characteristics of the separate types of junction transistors, the batteries all have the proper polarities for correct operation of this direct-coupled arrangement. This would not be possible with transistors all of the same type, when high-gain grounded-emitter circuits are employed in the stages.

Fig. 412 shows a direct-coupled amplifier which allows use of identical transistors in each stage. The arrangement is a grounded

emitter followed by a grounded base. P-n-p junctions are assumed. Battery B1 has correct polarity for supplying the two emitters, and B2 for the two collectors. This system has lower overall gain than the one shown in Fig. 411 because of the reduced gain of the grounded-base stage.

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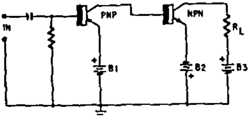

Fig 411 Direct-coupled amplifier using p-n-p and n-p-n transistors

frequencies, provided the transistors themselves will reach the frequencies of interest.

Many factors limit transistor frequency range. One is the diffusion time of electrons and holes in the semiconductor. Reactive

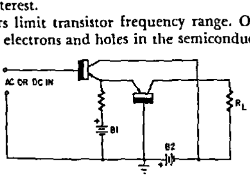

Fig 412. Direct-coupled amplifier using identical transistor types

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Transistor rf amplifiers may be operated single-stage, single-ended, push-pull and in cascade similar to audio amplifiers. Interstage coupling can be capacitive, inductive, transformer type, and in some cases direct.

At radio frequencies, as at dc and low frequencies, the input impedance of the transistor is lower than the output impedance (except in the case of the junction type transistor used in a grounded-collector circuit). The same requirement for a step-down interstage impedance ratio thus holds at radio frequencies. Highest rf power gain is obtained through transformer coupling between stages and with transformers at the input and output of a multistage amplifier.

Fig. 413 shows the circuit of a grounded-emitter. transformer-coupled amplifier using a high-frequency transistor. The input

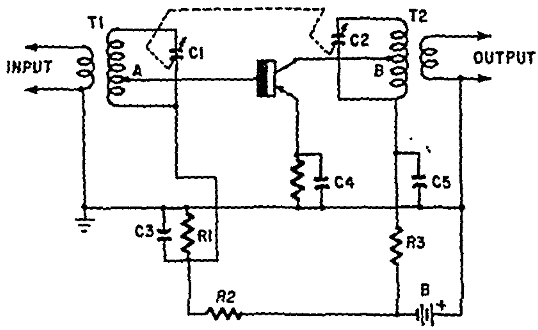

Fig. 413. *Single-stage transformer-coupled rf amplifier.*

and output transformers are tuned simultaneously by the two-gang capacitor, C1-C2. For impedance matching, the base of the transistor is tapped down on the secondary (A) of the input transformer T1 and the collector on the primary (B) of the output transformer T2. A single battery is employed and dc base bias is obtained though voltage divider R1-R2.

Fig. 414 is another version of the transformer-coupled amplifier, shunt dc feed being used in this circuit. Here, R1 is the constant-current emitter resistor. R2 and R3 represent a grounded center-tapped bleeder across the battery or power supply, a simple method for getting positive and negative voltages from a single source. C5 and C6 are filter capacitors. Collector current is fed through a radio-frequency choke, RFC, the inductance of which can be a stock value between 1 and 85 millihenries, the lower values being used for the higher radio frequencies.

Impedance coupling is employed in untuned and wide-band amplifiers with somewhat better success than straight R-C coupling. An example is shown in Fig. 415. The coupling inductor L is connected in the collector circuit only and can be a high-Q rf choke or a specially designed coil. An inductor also might be con-

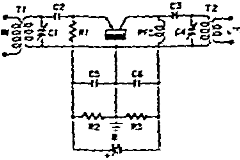

Fig. 414. Transformer-coupled amplifier with shunt feed.

nected in series with the current-stabilizing resistor R and the junction of C1 and the emitter.

When using point-contact transistors, parallel resonant tuned circuits, such as indicated in Figs. 413 and 414, occasionally encourage instability if the transistor tends toward high α and r_b. The reason is that the impedance of the parallel-resonant combination drops to low values at frequencies removed from resonance. This might encourage oscillation if the transistor is short-circuit unstable, although the tuned-circuit impedance is high at its resonant frequency. This difficulty is minimized in the circuit shown in Fig. 416 by the use of a series-resonant coupling circuit. The tuned coupling circuit. L–C2. passes maximum current at resonance and works directly into the current-actuated emitter of the second transistor The R-C-coupled output of the second stage may be supplanted by another series L-C arrangement for coupling to a third stage.

An extension of this idea is the double-tuned amplifier of Fig. 417 which has tunable series-resonant circuits in both emitter and collector of cascaded stages. The collector circuit of the first stage is tuned by L1–C2, and the emitter circuit of the second stage by L2-C4. Capacitor C3 is for interstage coupling. Coils L1 and L2 are shielded or isolated from each other to minimize magnetic coupling. Bias is shunt-fed to the emitter and collector through resistors R1 and R2. The blocking capacitors, C1 and C5, are large with respect to tuning capacitors C2 and C4.

fiers and similar channels of a fixed-tuned nature. In such cases, capacitors C2, C3, and C4 are screwdriver-adjusted trimmers. This circuit usually is not applied to continuously-tuned rf amplifiers because of the prohibitive number of tuning-capacitor gangs required for a multistage arrangement.

Each of the amplifiers shown in Figs. 413 to 417 may be cascaded with others of the same type or of the other types shown, for increased gain and, except in the case of Fig. 415, for increased selectivity.

Amplifier characteristics

Current, voltage and power amplification all are afforded by transistor circuits. Complex formulas for power amplification and for transistor resistances may be found in Chapter 3.

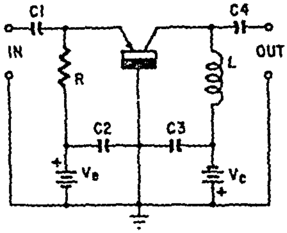

Fig. 415. *Impedance-coupled rf amplifier.*

Table 4—2 gives *simplified* formulas for calculating current, voltage and power amplification, as well as input and output impedances of transistor amplifiers of all three configurations. The symbols in these formulas are the same as those defined in preceding chapters.

Transistor bias stabilization

When a grounded-emitter transistor is operated from a single dc supply, the bias current flowing in the base-to-emitter circuit which determines the operating point of the amplifier sometimes is obtained through a large series resistance (R1 in Fig. 418-a) connected between the base and the supply. However, transistor operation is not stable with this arrangement, the reason being that most of the emitter current results from the flow of current from the collector to the emitter and not from the flow through R1. When i_c increases with temperature, as a result of i_{co} drift, i_e also increases. Because of the inherent regenerativeness of the grounded emitter, i_c increases still further—the process continuing in a *runaway* until the transistor is damaged. The series-re-

Table 4—2. Approximate Transistor Amplifier Formulas

Transistor Amplifier Connection	Input Impedance	Output Impedance	Current Amplification	Voltage Amplification	Power Gain	Vacuum-Tube Analogy
Grounded-Emitter	$\frac{r_e + r_b(1-a)}{1-a}$	$r_c(1-a) + r_e$	$\frac{-a}{1-a}$	$\frac{-a R_L}{r_e + r_b(1-a)}$	$\frac{a^2 R_L}{[r_e + r_b(1-a)](1-a)}$	Corresponds to grounded-cathode amplifier
Grounded-Base	$r_e + r_b(1-a)$	r_c	a	$\frac{a R_L}{r_e + r_b(1-a)}$	$\frac{a^2 R_L}{r_c + r_b(1-a)}$	Corresponds to grounded-grid amplifier
Grounded-Collector	$\frac{r_b + R_L}{1-a}$	$r_e + R_G(1-a)$	$\frac{1}{1-a}$	1	$\frac{1}{(1-a)}$	Corresponds to grounded-plate amplifier (cathode follower)

Where $a = \frac{r_m}{r_c} \cong \left.\frac{\Delta I_c}{\Delta I_e}\right]_{V_c = \propto}$

Table, Courtesy Texas Instruments, Inc.

The double-tuned amplifier is suited particularly to if amplifiers and similar channels of a fixed-tuned nature. In such cases, capacitors C2, C3, and C4 are screwdriver-adjusted trimmers. This circuit usually is not applied to continuously-tuned rf amplifiers because of the prohibitive number of tuning-capacitor gangs required for a multistage arrangement.

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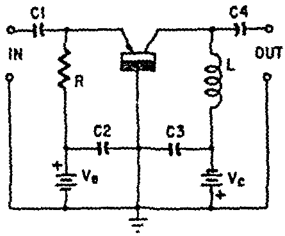

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Table 4–2 gives *simplified* formulas for calculating current, voltage and power amplification, as well as input and output impedances of transistor amplifiers of all three configurations. The symbols in these formulas are the same as those defined in preceding chapters.

Transistor bias stabilization

When a grounded-emitter transistor is operated from a single dc supply, the bias current flowing in the base-to-emitter circuit which determines the operating point of the amplifier sometimes is obtained through a large series resistance (R1 in Fig. 418-a) connected between the base and the supply. However, transistor operation is not stable with this arrangement, the reason being that most of the emitter current results from the flow of current from the collector to the emitter and not from the flow through R1. When i_c increases with temperature, as a result of i_{co} drift, i_e also increases. Because of the inherent regenerativeness of the grounded emitter, i_c increases still further—the process continuing in a *runaway* until the transistor is damaged. The series-re-

Table 4—2. Approximate Transistor Amplifier Formulas

Transistor Amplifier Connection	Input Impedance	Output Impedance	Current Amplification	Voltage Amplification	Power Gain	Vacuum-Tube Analogy
Grounded-Emitter	$\frac{r_e + r_b(1-a)}{1-a}$	$r_c(1-a) + r_e$	$\frac{-a}{1-a}$	$\frac{-a R_L}{r_e + r_b(1-a)}$	$\frac{a^2 R_L}{[r_e + r_b(1-a)](1-a)}$	Corresponds to grounded-cathode amplifier
Grounded-Base	$r_e + r_b(1-a)$	r_c	a	$\frac{a R_L}{r_e + r_b(1-a)}$	$\frac{a^2 R_L}{r_e + r_b(1-a)}$	Corresponds to grounded-grid amplifier
Grounded-Collector	$\frac{r_b + R_L}{1-a}$	$r_e + R_G(1-a)$	$\frac{1}{1-a}$	1	$\frac{1}{(1-a)}$	Corresponds to grounded-plate amplifier (cathode follower)

$$\text{Where } a = \frac{r_m}{r_c} \cong \left.\frac{\Delta I_c}{\Delta I_e}\right]_{V_c = \alpha}$$

Table, Courtesy Texas Instruments, Inc.

as method shown in Fig. 418-a and appearing
is satisfactory only where the ambient temperature is con-
nd the transistor is operated so well within its ratings that
al heating is not appreciable. An additional shortcoming
circuit is that transistors cannot be interchanged unless they
osely matched.

Fig. 418-b, the dc base bias is obtained from a voltage di-

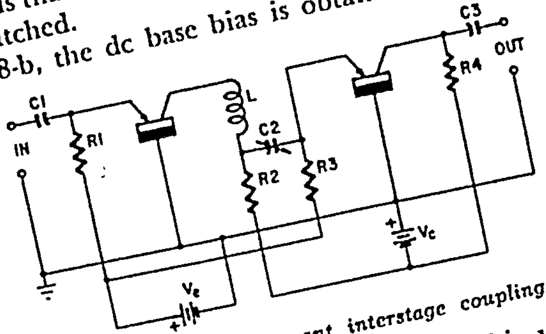


Fig. 416. *Series-resonant interstage coupling.*

vider, R1-R2, operated from the supply. If this voltage divider is designed for good regulation, the base-to-emitter current re-

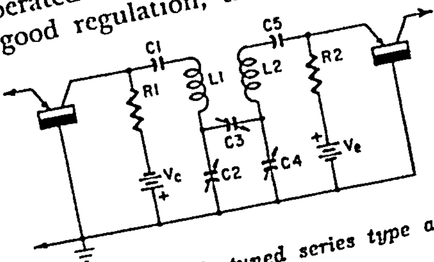
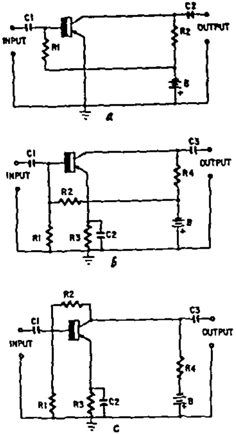


Fig. 417. *Double-tuned series type amplifier.*

mains substantially constant regardless of collector-current excur-
sions. An emitter resistance, R3, analogous to the cathode resistor
in a tube circuit, determines the value of emitter current which
will flow for a given dc base voltage at the junction of R1 and R2.
For good voltage regulation, R1 must be low and the curre
through the voltage divider high. However, the divider resista
must not be so low as to load the input signal source severely
to draw excessive current from the supply. A convenient rule
thumb is to proportion the divider for a base current of 10%
the selected collector current. The divider impresses a con
voltage (v_b) on the base. The emitter voltage (v_e) is substan
equal to v_b. The emitter current (i_e) may then be set to a d
value by selection of the emitter resistance, R3. The requi

sistance of R3 is determined by means of Ohm's law from v_b and the desired collector current i_c: $R3 = v_b/i_c$. Resistor R3 then is bypassed with capacitor C2 to prevent loss of amplification due to degeneration.

This is a stabilized bias arrangement with respect both to



Figs. 418-a, -b, -c. *Various methods of biasing transistors.*

temperature and to transistor interchangeability. In practice, R1 *has a resistance between 2,000 and 20,000 ohms for convention*al, low-powered transistors, and the input impedance of the stage runs around 1,000 ohms. The value of R2 depends upon the dc supply voltage and the desired voltage division.

Fig. 418-c shows another, similar, bias arrangement. Unlike the preceding example, however, the base voltage divider is connected between the collector and ground. This scheme works best with high supply voltages and in the absence of wide variations in

sistor bias method shown in Fig. 418-a and appearing in many circuits is satisfactory only where the ambient temperature is constant and the transistor is operated so well within its ratings that internal heating is not appreciable. An additional shortcoming of this circuit is that transistors cannot be interchanged unless they are closely matched.

In Fig. 418-b, the dc base bias is obtained from a voltage di-

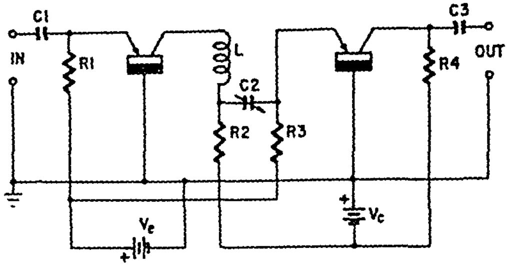


Fig. 416. *Series-resonant interstage coupling.*

vider, R1-R2, operated from the supply. If this voltage divider is designed for good regulation, the base-to-emitter current re-

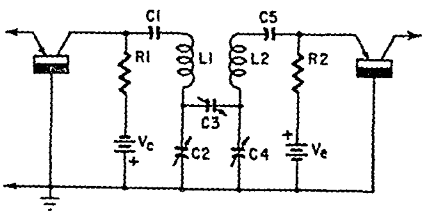


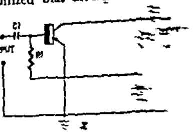
Fig. 417. *Double-tuned series type amplifier.*

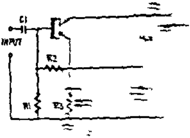
mains substantially constant regardless of collector-current excursions. An emitter resistance, R3, analogous to the cathode resistor in a tube circuit, determines the value of emitter current which will flow for a given dc base voltage at the junction of R1 and R2.

For good voltage regulation, R1 must be low and the current through the voltage divider high. However, the divider resistance must not be so low as to load the input signal source severely nor to draw excessive current from the supply. A convenient rule of thumb is to proportion the divider for a base current of 10% of the selected collector current. The divider impresses a constant voltage (v_b) on the base. The emitter voltage (v_e) is substantially equal to v_b. The emitter current (i_e) may then be set to a desired value by selection of the emitter resistance, R3. The required re-

tance of R3 is determined by means of Ohm's ... e desired collector current i_c: $R3 =$ ...
passed with capacitor C2 to prevent loss ...
egeneration.

This is a stabilized bias arrang...





[illegible]

temperature and ...
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Fig. 415-c shows ...
the preceding ...
nected between the ...
with high supply voltage and ...

temperature. Transistors must be closely matched to allow interchangeability in this circuit.

These are the biasing circuits most widely used. There are many others achieving stabilization, some utilizing the temperature coefficient of semiconductor diodes and others employing thermistors.

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ransistor oscillators

MOST transistors oscillate readily. We learned in Chapter 2 that precautions often are necessary to *prevent* them from scillating. Their small size, light weight and low power requirenents suit transistors to many oscillator applications.

Any transistor amplifier can be made to oscillate by feeding portion of its output energy back to the input circuit in proper hase. In the grounded-base circuit, where there is no phase eversal through the transistor, this can often be done by means of a capacitor connected between the output and input. In the grounded-emitter circuit, inductive feedback coupling may be mployed with the transformer windings poled for positive feedback. In the grounded-emitter circuit using a junction transistor, frequency-selective feedback may be obtained through a phase-shifting R-C network.

The negative resistance characteristics of point-contact transistors also may be utilized for oscillation and pulse generation in simple R-C circuits. With the high-alpha point-contact transistor, positive feedback can be developed across a tuned circuit connected in the base-to-ground lead.

This chapter is devoted to practical circuit arrangements for af and rf oscillators. The circuit constants correspond to those used in amplifier operation of the same transistor. Iron-cored coils and transformers are employed for audio operation of these circuits, and air-wound units for rf operation. It is important to note here that most transistors will oscillate at frequencies higher than alpha cutoff. High-frequency oscillation is aided by

high collector voltages, as long as the maximum rated collector dissipation of the transistor is not exceeded.

In each of the circuits in this chapter, except Fig. 507, the

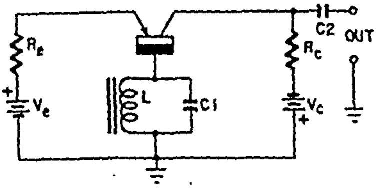


Fig. 501. *Base-tuned oscillator.*

bias supply polarities and transistor symbols are for either p-n-p junction or point-contact units. Opposite polarities for both V_e and V_c apply to the n-p-n junction transistor.

Base-tuned circuit

Fig. 501 shows an oscillator in which the frequency-determining tuned circuit, L–C1, is connected between ground and the base of the transistor. Since this circuit depends upon high alpha and to some extent upon emitter negative resistance for its operation, it is satisfactory for use only with point-contact transistors.

The collector supply voltage (V_c) and collector load resistance (R_c) have the values which ordinarily would be used for ounded-base amplifier operation of the transistor selected. Values of the emitter supply (V_e) and the emitter current-limiting resistor (R_e) must be chosen for rated emitter current, and R_e must be adjusted carefully for sine-wave output.

Capacitance-coupled collector output is shown. For maximum signal, the impedance of the load device should be at least 10 times the transistor output impedance. Output also may be obtained by inductive coupling to the coil, L. For this purpose, L might be one winding of a transformer and the output coil a second winding. Inductive coupling will allow maximum power output. Where only signal voltage with negligible power is required, the collector output connection should be used. This connection has the least detuning effect.

Where practicable, some improvement in operation can be had by tapping the base connection down on coil L for a closer base-impedance match. At high frequencies, it will be necessary to bypass both V_e and V_c.

Separate bias supplies are shown, but the two voltages can be

transformer T and this winding can be proportioned for proper impedance match to the load.

If the dc resistance of the transformer windings is high, voltages V_e and V_c of the two bias supplies must be increased proportionately to obtain rated transistor electrode voltages. Emitter current must be adjusted closely by variation of a series resistor to obtain sine-wave signal output. In Fig. 502-b, the base resistor R may be adjusted for this same purpose. When V_e or R is too low, peak clipping will take place and in extreme cases the circuit will behave like a blocking oscillator. The base may be biased also through a voltage divider, using the high-stability method described under *Transistor bias stabilization* in the preceding chapter.

Series-resonant feedback

In point-contact transistor circuits, there is some objection to the conventional inductive feedback arrangements shown in Fig. 502, although these circuits will work with most transistors. The reason for this is the debatable practice of shunting the high-impedance voltage-supplying parallel-resonant circuit with the low-impedance collector or emitter.

A more logical approach would appear to be use of a low-impedance current-supplying series-resonant circuit in the feedback path. Fig. 503 shows an oscillator employing this type of feedback. At resonance, the tuned feedback loop, L–C1, has its lowest impedance, hence highest feedback current is supplied to the emitter. This type of feedback can be used only in the grounded-base circuit when applied in the manner shown in Fig. 503. Point-contact and junction transistors both are satisfactory.

The phase of the transmitted signal is rotated 180° by the single-stage grounded-emitter circuit, so two cascaded grounded-emitter stages would be needed to obtain feedback voltage of the correct phase for oscillation. In most instances, the added expense and complication fail to justify the extra transistor stage.

The oscillation frequency is determined principally by the values of L and C1, although emitter and collector currents have strong influence. The values of V_e, V_c, R_c, and R_e are the same as required for amplifier operation with the chosen transistor. Emitter current must be adjusted, by variation of R_e, V_e, or both, for sine-wave output.

Two dc bias sources are shown, but a single source may be employed and the emitter and collector voltages taken from taps properly spaced on the center-grounded voltage divider. The

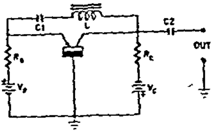
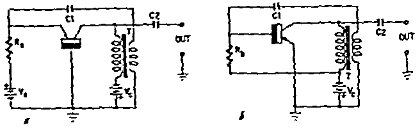


Fig. 503. *Series-resonant feedback oscillator.*

emitter supply, V_e, may be omitted entirely. However, when this is done, peak clipping and resulting high distortion of the



Figs. 504-a, b. *Transformer feedback in grounded-base and grounded-emitter circuits with series-tuned secondary.*

output signal occur at lower output levels than when fixed bias is provided.

Capacitively coupled high-impedance output is shown and is the most practical for this type of oscillator. Light loading will have negligible detuning and distorting effect

Inductively coupled series-resonant feedback

The circuits in Fig. 504 allow some of the advantages of both inductive output-input coupling and series-resonant tuning to be secured. Transformer T provides inductive coupling from the collector output. The secondary winding of this transformer forms a series-resonant circuit with the capacitor C1 connected back to the input. The oscillation frequency is determined principally by the secondary inductance and the capacitance of C1. The transformer must have a stepdown ratio, from collector to input, for impedance matching in the chosen transistor.

The grounded-base circuit (Fig. 504-a) is suitable for either point-contact or junction transistors, but the grounded-emitter circuit (Fig. 504-b) is for junction types only. Capacitive high-impedance output coupling is shown, and with high-impedance external loads will have the least detuning effect. Output also may be taken from a third winding on the transformer and this winding can be proportioned for exact match to the external load impedance. This latter connection permits maximum power to be drawn from the oscillator but has the greatest detuning effect.

Low-distortion sine-wave output may be obtained by adjusting R_e, V_e, or both (Fig. 504-a) and by adjusting R_b in Fig. 504-b.

R-C type oscillators

The simplicity and compactness of resistance-capacitance type circuits is as attractive in transistor oscillators as in similar tube circuits. However, not every tube type R-C oscillator can be transistorized readily.

Fig. 505 shows three practical R-C type transistor oscillator circuits. The arrangements in Figs. 505-a and -b are for point-

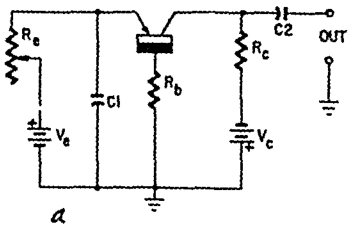


Fig. 505-a. *R-C transistor oscillator.*

contact transistors only, while the one in Fig. 505-c is for junction units only.

The circuit in Fig. 505-a is used frequently as a pulse generator or multivibrator. Switch-type point-contact transistors are especially suited to this circuit. With a high-alpha unit, appreciable positive feedback voltage is developed across the external base resistance, R_b. The oscillation frequency is determined by capacitor C1, resistance R_e, and the input resistance of the transistor. Emitter and collector currents and the value of R_b also influence the frequency at any fixed setting of C1 and R_e.

As the value of R_e is varied, the oscillation frequency will jump

abruptly from one discrete value to another, in the familiar manner of a multivibrator. With large values of feedback, the output waveform consists of steep unilateral pulses. At low feedback

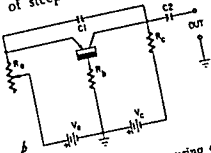


Fig 505-b. R-C oscillator using capacitive feedback.

levels, the output waveform becomes more rounded and bilateral, nearly sinusoidal at very low values.

The external collector resistance, R_c, has the same value i would have in an amplifier or switching circuit employing th

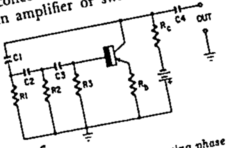


Fig 505-c R-C oscillator using phase-shifting network

same transistor. The base resistor, R_b, will be between 1 2,000 ohms, depending upon the model and manufacture transistor.

The oscillator may be synchronized readily with a sign is capacitance-coupled between emitter and ground, or ground.

Capacitance-coupled high-impedance output is show may be taken also at lower impedance capacitively f base resistor R_b. A very high impedance load device to prevent detuning the oscillator or degrading its o form.

The circuit shown in Fig. 505-b is a simple groun plifier with external base resistance R_b. Positive feedl

oped across R_b. Capacitor C1 also provides capacitive feedback coupling between collector and emitter. In some instances, R_b may be omitted—the base being grounded directly. Like the circuit shown in Fig. 505-a, this arrangement is satisfactory only for use with point-contact transistors.

The oscillation frequency is determined principally by the values of C1, R_e, and the input resistance of the transistor, although it is sensitive also to emitter and collector current levels and to the value of R_b. For any given setting of the resistances and currents, the frequency is inversely proportional to the value of C1.

The external collector resistance, R_c, is the same value that would be used in a grounded-base amplifier or switching circuit employing the same transistor. Emitter current may be set, by adjusting R_e or V_e, for shaping of the output wave. The output waveform will vary from steep unilateral pulses for heavy feedback to nearly sinusoidal for light feedback.

Capacitively coupled high-impedance output is shown, but lower-impedance output can be obtained capacitively across the base resistor, R_b. High-impedance load devices must be employed to prevent detuning and waveform deformation.

A phase-shift oscillator circuit is shown in Fig. 505-c. Here, a grounded-emitter arrangement is used with a junction transistor. The grounded emitter resistor, R_b, provides stabilization through degeneration.

The feedback network is composed of three cascaded sections (C1–R1, C2–R2, and C3–R3) which provide 180° of total phase shift between the collector output and base input. Each section provides 60° shift. The problem of a phase-shift network for a transistor is somewhat more complicated than one for a vacuum tube, chiefly because of the low input impedance of the transistor. The last resistance in the R-C combination is not simply R3, but the transistor input resistance in parallel with R3. In terms of the resistance of each leg, the capacitance required in that leg may be found from the relationship: $C = 0.289/3.14R$.

Resistor R_b should be kept as low as possible, since it reduces the gain of the transistor. A high output signal voltage is needed at the collector to obtain enough signal at the base of the transistor, after successive divisions in the phase-shift network, to sustain oscillation. For this reason, high amplification is required.

The output is high-impedance, and a high-impedance load device must be used for minimum loading, detuning, and waveform

deformation. Stability and positive action of the oscillator may b improved by using three cascaded grounded-emitter amplifie stages in place of the single stage shown. This will provide highe amplification before feedback. The odd number of stages is re quired for proper phasing.

Under proper operating conditions of base and collector cur

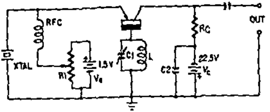


Fig 506. *Crystal at emitter input.*

rents, good output waveform can be obtained. Resistance R_b may be adjusted for the best compromise between quick starting, good waveform, and output signal amplitude.

Crystal oscillators

Quartz crystals may be connected into transistor oscillators ir a variety of ways. Fig. 506 shows one method in which the crysta is connected between emitter and base of a point-contact tran sistor, and the tuned circuit, L–C1, operated between base anc

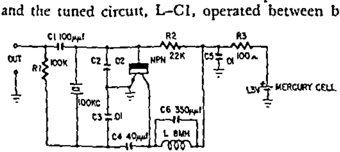


Fig 507 *Crystal-controlled 100-kc standard-frequency oscillator*

ground. This circuit is covered by U.S. Patent No. 2,570,436 issued to Everett Eberhard and Richard O. Endres and assigned to Radio Corporation of America.

Emitter bias is adjusted by means of potentiometer R1. The parallel resonant circuit, L–C1, is tuned to the crystal frequency. The external collector resistance, R_c, is adjusted for normal collector current with the 22.5-volt bias source, V_c.

Capacitively coupled high-impedance output is shown. Outpu

also may be obtained by inductive coupling to coil L. This type of coupling will allow the most power to be drawn from the circuit.

Fig. 507 shows a crystal-controlled 100-kc standard-frequency oscillator developed at the National Bureau of Standards. This unit, employing an n-p-n junction transistor, operates with such economy that it can be left running continuously. This is a decided advantage with standard frequency oscillators since continuous operation eliminates frequency drift and lost time due to warmup periods after switching on. The oscillator draws 100 microamperes dc from the single 1.35-volt mercury cell and is estimated to run *continuously* for 5 or more years before the battery must be replaced.

Tests at the Bureau of Standards show this oscillator to have a long-term drift of only 3 parts in 10^9 per day at 100 kc. Its short-term stability is 3 parts in 10^{10}. The frequency varies approximately 1 part in 10^8 per degree Centigrade temperature change and 1 part in 10^8 per 0.10-volt supply-voltage variation.

The n-p-n transistor is used in the grounded-emitter connection. A capacitive voltage divider (C3 and C4, in series) reduces the rf signal voltage from the tuned circuit (L–C6) before it is applied to the crystal. An output of 0.8 volt is provided by the oscillator.

High-frequency oscillator

The upper frequency limit of an oscillator using a conventional transistor follows somewhat the frequency characteristics of the transistor as an amplifier. However, it is well known that transistors will oscillate at frequencies beyond alpha cutoff, especially if higher collector voltages are used without exceeding maximum allowable collector dissipation.

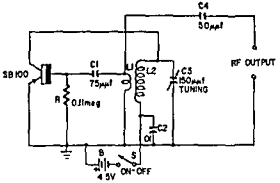
Fig. 508 shows the circuit of a high-frequency oscillator employing an SB100 surface-barrier transistor. Tuning of this circuit is continuously variable, by means of C3, from approximately 15 to 44 mc. Higher-frequency operation may be obtained with individual transistors by reducing the inductance of coils L1 and L2.

This is a simple, tickler type oscillator. When the polarity of L2 is correct, positive feedback is provided through the inductive coupling between collector and base. Total dc drain of the oscillator is approximately 1.1 ma from the 4.5-volt battery.

Oscillator tuning

The frequency of a transistor oscillator, like that of a vacuum-tube circuit of the same kind, can be varied continuously or in

steps by varying or switching one or more of the frequency-determining circuit constants. Thus, in Figs. 501, 502, 503, and 504, the frequency-determining capacitor may be switched in value in audio oscillators, or a variable capacitor can be used at radio fre-



L1: 7 turns No. 20 enameled wire closewound adjacent to ground end of L2.

L2: 9 turns No 20 enameled wire on ½-inch-diameter form Space to winding length of ½ inch.

Fig 508 *High-frequency oscillator using a surface-barrier transistor.*

quencies. The associated coil or transformer winding may be switched in value to change ranges when the capacitor is variable.

It will be more convenient in Figs. 505-a and -b to vary or change the value of resistor R_x to change frequency, and to change or switch capacitor C1 to change range. In the phase-shift oscillator (Fig 505-c) all three network capacitors (C1, C2, and C3) must be varied simultaneously, or all three network resistors (R1, R2, and R3) simultaneously, to change frequency

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11. Lawrence Fleming, "Broadcast-Band Test Oscillator Using Transistors," *Radio & Television News,* October, 1955; p. 48.

duality in transistor circuit design

IN electrical engineering, one component or circuit is said be the *dual* of another component or circuit when current one behaves like voltage in the other. Thus, inductance is the du of capacitance, current of voltage, parallel of series, impedance admittance, open circuit of short circuit, etc. The reverse al is true. For example, if resistance is the dual of conductance, th conductance is the dual of resistance. Very important to t present study, the transistor is considered to be the dual of t vacuum tube.

Table 6—1 lists several common components, properties, a circuits with their duals The table may be read across in bo directions. Thus, line 8 is read either "node is the dual of loo or "loop is the dual of node."

On the basis of current and voltage behavior, many duals electronic circuitry will suggest themselves to the reader: A seri resonant circuit is the dual of a parallel-resonant circuit, a shu bypass capacitor is the dual of a series choke coil, and the dual a current through a capacitor is a voltage across a coil

Figs. 601-a and -b show some simple circuit elements and c cuits with their duals. The dual of a voltage-stepup transform is a current stepdown transformer. In Fig. 601-c note that ea series capacitor in the T section is replaced by a shunt induct in the pi-section dual, and the shunt inductor in the T section a series capacitor in the pi section. Coupled coils in a circuit w mutual inductance present an interesting case illustrated by t transformer in Fig. 601-d. The transformer is represented by

inductive T network when its mutual inductance M is less than either the primary inductance (L_p) or the secondary inductance

Table 6–1. Common Components, Properties, and Circuits With Their Duals

Item	Dual
1. Resistance (R)	Conductance (G)
2. Inductance (L)	Capacitance (C)
3. Impedance (Z)	Admittance (Y)
4. Voltage (E, V)	Current (I)
5. Tube	Transistor
6. Plate current (I_p)	Collector voltage (v_c)
7. Plate voltage (E_p)	Collector current (i_c)
8. Node	Loop
9. Series	Parallel
10. Stepup transformer	Stepdown transformer
11. Parallel capacitance	Series inductance
12. Inductive T	Capacitive pi
13. Voltage supply	Current supply

(L_s). The dual of this inductive T is the capacitive pi. (See item 12 in Table 6–1.)

Applications to transistor circuit design

Duality is a useful tool for converting well-known vacuum-tube circuits into transistor circuits. Briefly, the technique is to put circuit elements w[illegible]h current characteristics in place of the tube-circuit elemen[illegible] have corresponding voltage character-istics. While [illegible]ation can be handled completely by duality, th[illegible]able to a number of transformations from tu[illegible]r circuits.

In F[illegible] duality between transistor and tube chara[illegible] The families of transistor collector voltag[illegible]r correspond to several constant emit-ter o[illegible]s, while the dual plate current vs. plate vol[illegible]corresponds to several constant grid volt-age[illegible]een.

A simple [illegible]ing of a tube circuit with a transistor in place of a tube and with substitutions of voltage-operated circuit elements for current-operated elements, and vice versa, will give a pictorial

view of the new circuit configuration. The quantitati
determining circuit component values and exact config

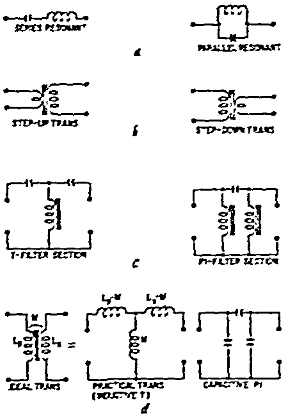

Fig. 601. Circuit elements with duals

not so simple, however. The procedure is to write fi
plete set of Kirchhoff's equations, expressing complete
rent and voltage relations in the tube circuit which is to
formed. Next, voltages are replaced with currents, an
with voltages in every part of the equations. Also, ea
must be superseded by the symbol of the dual quantity
the circuit is redrawn from the transformed equations,
configuration for use with the transistor. The transis
itself becomes the dual of the tube circuit and should
more efficiently than when the transistor simply is subst
a tube in a conventional tube circuit.

A convenient device for use in duality calculations

a given tube circuit and a desired transistor circuit is *transformation resistance* (r). The quantity r consists of two parts, r_1 and r_2, so related that $r_1 r_2 = r_p r_c$, where r_p is the tube plate resistance

Table 6—2. Duals in Terms of Transformation Resistance

Tube-Circuit Quantity	Transistor-Circuit Dual	Value of Dual
R	G	R/r_1r_2
i	e	ir_2
e	i	e/r_1
L	C	L/r_1r_2
Z	Y	Z/r_1r_2
e_p	$-i_c$	e_p/r_1
i_p	$-e_c$	i_pr_2
e_g	$-i_e$	e_g/r_1*
i_g	$-e_e$	i_gr_2*
r_p	r_c	r_1r_2/r_p
μ	α	μ
G_m	Υ_m	$G_mr_1r_2$

*For grounded-base connection

and r_c the transistor collector resistance. Both r_1 and r_2 may be chosen at will. As an illustration of the use of the transformation resistance, a series inductor L in the tube circuit is replaced in the transistor circuit by a shunt capacitor C with a value of L/r_1r_2. Table 6—2 shows several tube-circuit constants with corresponding transistor-circuit duals and the relationship of the latter to the former in terms of transformation resistance.

Examples of dual circuits

Fig. 602 shows three simple examples of common tube circuits and the corresponding transistor dual circuits. In Fig. 602-a, a double-tuned vacuum-tube amplifier (of the type commonly found in receiver if stages) has the plate of the first tube tuned

by parallel-resonant circuit L1–C1, and the grid of the second tube tuned by a similar parallel-resonant circuit, L2–C2. The two coils are inductively coupled by mutual inductance M. In the dual transistor circuit, the plate parallel-resonant circuit of the tube amplifier becomes the series-resonant circuit, L3–C3.

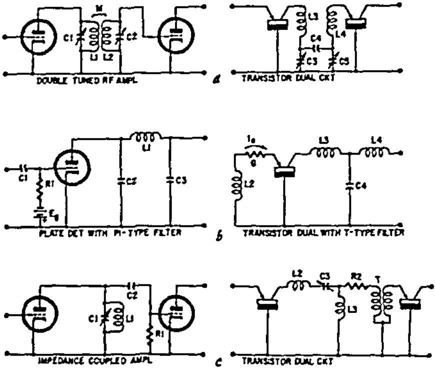

Fig. 602. Various tube circuits and their transistor duals.

Likewise, the grid parallel-resonant circuit in the tube amplifier becomes the series-resonant circuit L4–C5 in the transistor amplifier. Mutual inductance M is replaced by the capacitive coupling C4.

A plate detector is shown in Fig. 602-b. It is customary to operate a low-pass pi-section filter, such as L1–C2–C3, in the plate output circuit of such a detector. The series capacitor, C1, in the tube circuit becomes the shunt inductor, L2, in the transistor circuit. Shunt capacitors C2 and C3 become the series inductors, L3 and L4, and the series inductor, L1, becomes the shunt capacitor, C4. In the tube circuit, grid voltage E_g is applied to the tube through a series resistor, R1. This voltage source is across the grid input circuit of the tube. In the corresponding transistor

circuit, the voltage-supply-in-series-with-a-resistor is replaced with a current supply I_e, shunted by a conductance G. The curved arrow is the standard symbol for a current supply. Note that the transistor current supply, unlike the tube voltage supply, is in *series* with the input. In performing the transformation to dual· circuits, all voltage supplies in the tube circuit are replaced by current supplies in the corresponding transistor circuit.

Fig. 602-c shows a single-tuned impedance-coupled amplifier. This type of connection is found in the stages of some if strips and in the exciter and rf amplifier stages of radio transmitters. In the corresponding transistor circuit, L1–C1, the parallel-resonant circuit of the tube amplifier, is replaced by the series-resonant circuit, L2–C3. Series capacitor C2 becomes the shunt inductor, L3. Shunt resistor R1 becomes the series conductance, R2. T is an ideal phase-reversing transformer inserted here because the grounded-base transistor does not reverse phase. Its purpose is to provide the same phase shift introduced by the tube in the original circuit. It may be omitted when signal phase is of no consequence. In each example except the tube circuit in Fig. 602-b all dc supplies have been omitted for the sake of simplicity in the diagrams.

Practical limitations

Application of the principle of duality will, in many instances, yield a transistor circuit which will perform the functions of a tube circuit more efficiently than if the transistor and its bias supplies simply were substituted for the tube and its supplies in the original circuit. However, there are some cases where the best type of circuit will not be obtained by dualizing a known tube circuit, but only by designing a specific transistor circuit from the start.

The assumption of duality between the transistor and tube, except for the fact that the former is current-actuated and the latter voltage-actuated assumes that both devices have similar dissipations and that the mu of the tube equals alpha of the transistor. This cannot be true of modern tubes, since the alphas in contemporary transistors available commercially vary from 0.90 to 0.996 for junction types, and from 2 to 3 for point-contact types.

Specific functions of some components in the tube circuit must be considered before attempting to dualize these components in the transistor circuit. For example, a series capacitor might have been included between stages in a tube circuit solely for the

Recommended Reading

1. R. L. Wallace and G. Raisbeck, "Duality as a Guide in Transistor Circuit Design," *Bell System Technical Journal*, April, 1951; p. 381.

2. Gordon Raisbeck, "Transistor Circuit Design," *Electronics*, December, 1951; p. 128.

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triggers and switches

TRIGGER and switching circuits differ from amplifiers in that the output of these circuits is not a continuous reproduction of an input signal, as it would be in a good amplifier. As the term *switch* implies, the circuit is either on or off—that is, output is either present or absent, low or high. The action is comparable to that of a relay or a mechanical switch.

The control signal, called the *trigger or pulse,* snaps the circuit into full-output condition or back to low- or zero-output condition without stopping at any intermediate point. The control-signal amplitude being smaller than the output amplitude, the switching circuit has power sensitivity. Electronic switching circuits, of which the vacuum-tube flip-flop is a familiar example, are highly desirable because they can switch back and forth at speeds greatly exceeding those possible with the fastest electromechanical devices, like relays. In high-speed computers, for example, electronic switching circuits often operate at a 1-mc rate (switching time of 1 microsecond).

The transistor offers considerable attraction as a switching device, particularly in complicated machines like counters and computers where many such circuits are needed, because of their small size, low power requirements, cool operation, and long life. As in other areas of transistor application, however, numerous problems such as uniformity of characteristics, drift, temperature dependence, transit time, etc., are to be solved before large-scale use will be practicable. The transistor switching circuit, like its vacuum-tube counterpart, offers noteworthy improvements over electromechanical switching devices

A considerable amount of work on transistor switching circuits already has been done in many laboratories, and much experience has been gained. Practical circuits resulting from this research and development take many forms. The broad subject of transistor switching circuits has so many ramifications that its full, detailed coverage is beyond the scope of this book. This chapter describes the basic action of these switching circuits.

Switching action through transistor negative resistance

In continuation of the discussion introduced under *Negative resistance* in Chapter 2, Fig. 701 shows an n-type point-contact transistor connected in a grounded-base circuit to display emitter

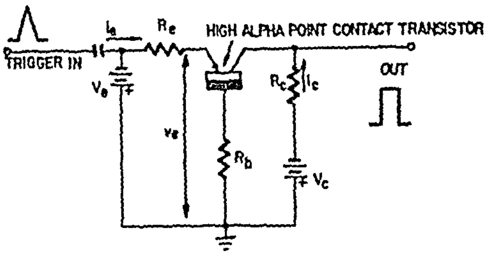


Fig. 701. Basic switching circuit using emitter negative resistance.

negative-resistance characteristics. The current amplification factor, alpha, of the transistor must be higher than 1, hence the restriction to the point-contact type.

External resistors, R_e, R_b, and R_c, are connected to the emitter, base, and collector electrodes, respectively. When the emitter current, i_e, is varied continuously, emitter voltage v_e, describes a plot such as shown in Fig. 702. The region A–B, when emitter current is negative, shows a small current excursion for a rather large voltage change and for obvious reasons is termed *cutoff*. This is not a true cutoff condition in the sense that we are accustomed to use the term in vacuum-tube practice, but one representing a current flow small enough to justify the term. Region A–B represents a positive resistance. The voltage change in region B–C is reversed with respect to the direction of emitter-current change, hence denotes negative resistance. The third region, C–D, in which the emitter voltage change is slight for a rather large increase in emitter current, is termed *saturation* and represents a positive resistance.

Thus, there are three discrete regions—A–B, B–C and C–D—in

the v_e vs. i_e characteristic, with an *upper turning point* (B), *lowe* *turning point* (C), and two positive-resistance regions separate by a negative-resistance region. Although the entire characteristi is nonlinear, each of its three regions can be considered separatel as linear for purposes of analysis.

The transistor will be unstable when the values of R_e and

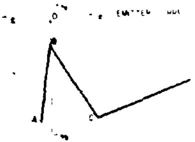


Fig. 702. Emitter negative-resistance curve.

are such that the operating point is within the negative-resistanc region B–C. But it will be stable when the operating point is i either the A–B or C–D region.

Basic bistable switching circuit

With the proper bias voltages, the value of external emitt resistance, R_e, may be chosen to place the operating point in ar of the three regions of the characteristic curve. The requiremen

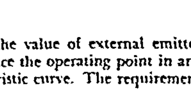
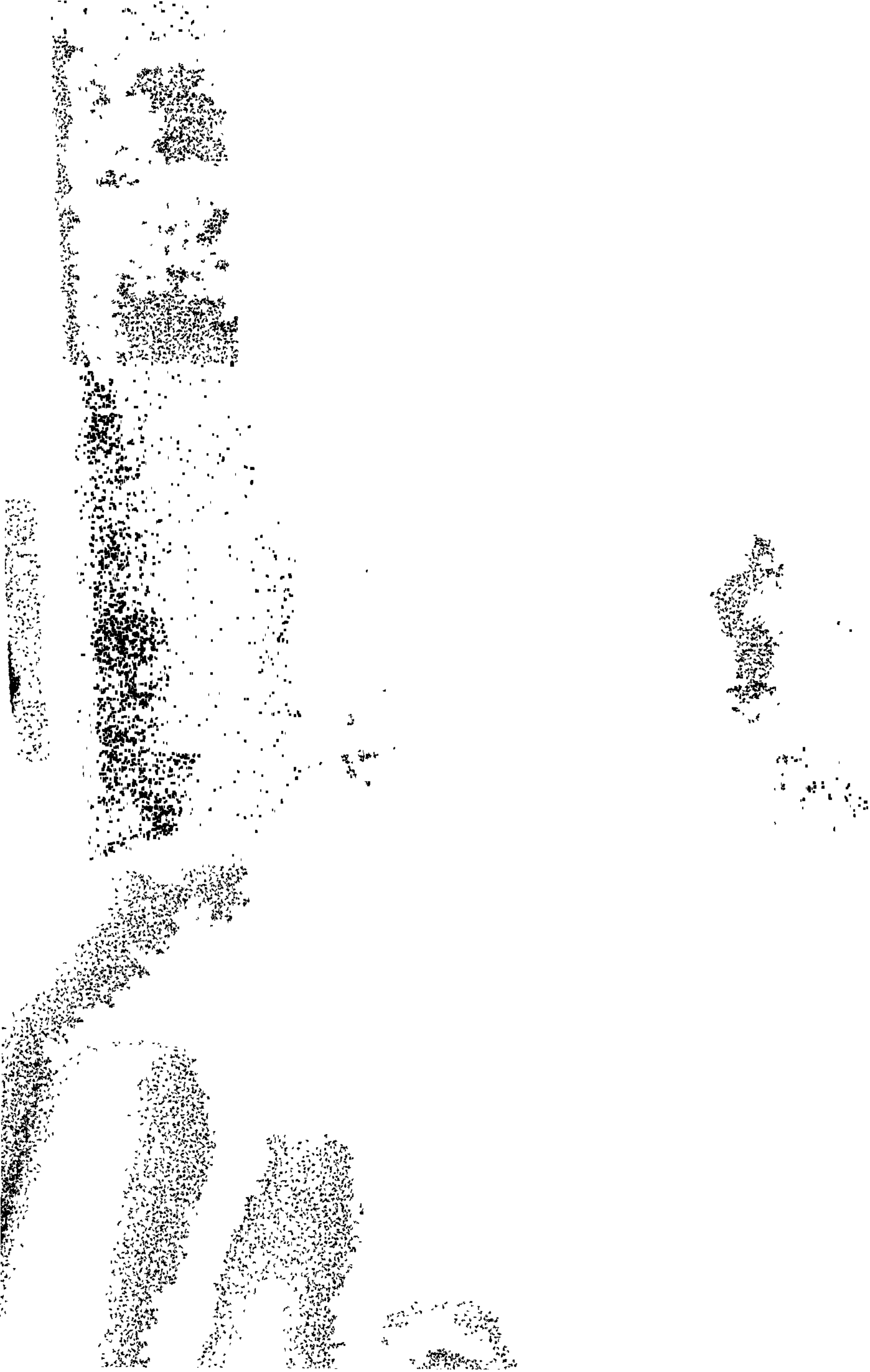


Fig. 703. Emitter negative-resistance characteristic with single operating points

that the load resistance be higher in value than the negative r sistance must be satisfied to achieve single operating points, shown in Fig. 703.

When the value of R_e is *less* than the negative resistance, illustrated by the dashed line in Fig. 704, and the bias voltag are adjusted properly, the load line can be made to intersect eac region once, and to give the multiple operating points *a*, *b*, and

The circuit may be adjusted for operation in the manner d



or flip-flop. The speed at which the operating point moves between the ON and OFF conditions is governed by operating parameters and by the internal properties of the transistor itself. Commercial switching type transistors are available with rise times as fast as 0.10 microsecond (10 mc). In general, fall times are somewhat slower than rise times.

Monostable circuit (single-shot) with point-contact transistor

A *single-shot* or *one-shot* circuit by successive ON–OFF action gives one complete output pulse for each input trigger pulse.

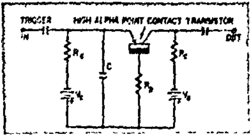


Fig. 705. *Single-shot (monostable) circuit.*

Note, however, that this is a true switching action and not merely *amplification of the control pulse.*

The simple transistor switching circuit already described can be adapted for single-shot operation by the addition of a capacitor C between emitter and ground, as shown in Fig. 705.

For an illustration of the monostable circuit characteristic, refer to Fig. 706-a. The values of *emitter external resistance* R_e and of negative emitter bias V_e are chosen such that R_e has a value less than the *negative-resistance* slope, B–C, and the load line intersects the characteristic at one point, *a*, in the cutoff region.

When the emitter voltage v_e is reduced, as by application of a positive pulse to the emitter electrode, the operating point is moved up the cutoff region and around the upper turning point, B. However, capacitor C cannot charge instantaneously (in fact, it resists a voltage change) and it effectively short-circuits the emitter to immediate voltage *changes.* The operating point accordingly does not move in the normal fashion over to D, but flips quickly to the second current value which i_e can have at voltage B. This point is C in the saturation region. But C in this case is not completely stable. Capacitor discharge brings the operating point slowly from C down to the lower turning point, D. This is another point only apparently stable, so the operating point flips quickly to the second current value at this voltage, which is point *e* in the cutoff region. From this point, also only apparently stable, the

pulse to the emitter) to move the operating point from a
ver the upper turning point, B, the operating point will
kly into the saturation region C–D and will fall back to c
, returns to point a. The operating point does not pause
ce operation is unstable in the B–C negative-resistance

rcuit now will operate indefinitely at point c which rep-
high collector-current level. This is the ON condition
vitching circuit. Further application of positive pulses
nitter will have no further switching effect. Note that
on resembles that of a thyratron tube in which all grid
s lost once the tube has been fired by the grid pulse.

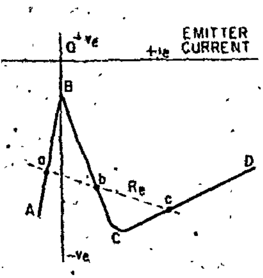


Fig. 704. *Emitter negative-resistance characteristic with multiple operating points.*

the circuit operating in a stable manner at point c, the
on of a negative trigger pulse of suitable amplitude to
ter will shift the operating point down and around the
rning point C. The operating point then flips to region
ce it cannot stop in the negative-resistance region B–C.
, pulses may be applied also to the base. A base pulse of
polarity will switch the circuit in the opposite direction.
the foregoing description, it is seen that operation is
uddenly from a to c by a positive emitter pulse, and
in from c to a by a negative pulse. The operating point
wherever it is until the next pulse of proper polarity
Collector output is low and the switch is said to be OFF
e operating point is at a. Conversely, collector output
nd the switch is said to be ON when the operating point

arrangement is a rudimentary, bistable transistor switch

or flip-flop. The speed at which the operating point moves between the ON and OFF conditions is governed by operating parameters and by the internal properties of the transistor itself. Commercial switching type transistors are available with rise times as fast as 0.10 microsecond (10 mc). In general, fall times are somewhat slower than rise times.

Monostable circuit (single-shot) with point-contact transistor

A *single-shot* or *one-shot* circuit by successive ON–OFF action gives one complete output pulse for each input trigger pulse.

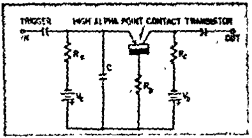


Fig. 705. *Single-shot (monostable) circuit*

Note, however, that this is a true switching action and not merely amplification of the control pulse.

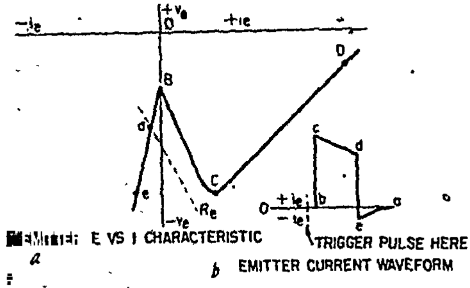
The simple transistor switching circuit already described can be adapted for single-shot operation by the addition of a capacitor C between emitter and ground, as shown in Fig. 705

For an illustration of the monostable circuit characteristic, refer to Fig. 706-a The values of emitter external resistance R_e and of negative emitter bias V_e are chosen such that R_e has a value less than the negative-resistance slope, B–C, and the load line intersects the characteristic at one point, *a*, in the cutoff region

When the emitter voltage v_e is reduced, as by application of a positive pulse to the emitter electrode, the operating point is moved up the cutoff region and around the upper turning point, B. However, capacitor C cannot charge instantaneously (in fact, it resists a voltage change) and it effectively short-circuits the emitter to immediate voltage *changes*. The operating point accordingly does not move in the normal fashion over to D, but flips quickly to the second current value which i_e can have at voltage B This point is C in the saturation region. But C in this case is not completely stable Capacitor discharge brings the operating point slowly from C down to the lower turning point, D. This is another point only apparently stable, so the operating point flips quickly to the second current value at this voltage, which is point *e* in the cutoff region. From this point, also only apparently stable, the

operating point then returns slowly to its original location, *a*. This finishes the complete cycle resulting from application of the trigger pulse.

Fig. 706-b shows the shape of the resulting emitter-current waveform. Points on this illustration have been lettered to correspond to those mentioned in the preceding explanation and ap-



Figs. 706-a, -b. *Monostable circuit characteristics.*

pearing along the curve in Fig. 706-a. The quick and slow changes may be detected easily in Fig. 706-b.

The circuit oscillates when R_e is higher than the negative resistance, and the load line intersects the curve at one point in the negative-resistance region. Oscillations are of the relaxation type, similar to those produced by corresponding gas-tube sawtooth generator circuits. Capacitor C charges through resistance R_e at a rate determined by the time constant of the R-C combination. It then is discharged by the transistor, and the cycle is repeated.

Monostable circuit with junction transistors

Fig. 707 shows a single-shot circuit employing two junction transistors. This is a monostable multivibrator similar to the comparable tube circuit. There is no output until a positive trigger pulse is applied to the input terminals. The circuit delivers one output pulse in response to each trigger pulse.

During the stable interval, transistor V2 is ON (conducting), since its base receives negative bias through resistor R4. The positive trigger pulse is applied through the coupling capacitor C2 and isolating diode D and causes V1 to conduct and V2 to cease conducting. At the end of the pulse, capacitor C2 begins to charge, increasing the base voltage of V2 toward the magnitude of the supply voltage. This action continues until the base voltage of V2

reaches a value equal to the emitter voltage. The emitter voltage, developed across the common emitter resistor R5 equals the supply voltage minus the base-to-emitter voltage drop across transistor V1. At this point, conduction is transferred abruptly to V2 and the circuit is restored to its original state.

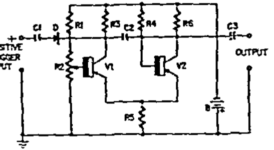


Fig. 707. Single-shot circuit using junction transistors.

The duration of the single-shot output pulse (pulse width) is adjustable by potentiometer R2, which controls the common-emitter voltage.

Bistable circuit (flip-flop) with junction transistors

The two-transistor flip-flop (Fig. 708) resembles the similar tube type Eccles-Jordan bistable multivibrator. Its operation also is similar.

The circuit is cross-coupled in the conventional manner, feedback being provided by the dc voltage dividers R2–R7 and R4–R5. The common-emitter resistor R6 biases the OFF transistor approximately to cutoff and, through its degenerative action, stabilizes the ON transistor. The transition of conduction from one transistor to the other is speeded by the shunting capacitors, C2 and C3, which couple the trigger pulse at one collector rapidly to the opposite base.

In the beginning when the battery is switched on, one transistor will conduct. If V1, for example, is conducting collector current, V2 is cut off because the voltage drop across the common emitter resistor R6 is sufficient for this purpose. Since V2 is not conducting, the base voltage of V1 (obtained through voltage divider R4–R5) is high since there is very little, if any, drop across R3. This maintains V1 in the conducting state.

Application of a negative trigger pulse to the input terminals momentarily raises the collector of V2 to a negative potential, initiating conduction in V2. The resulting decrease of V2 collector

voltage lowers the base voltage of V1 (through R4–R5) and therefore the collector current of V1. This action causes the collector voltage of V1 to rise toward the potential of the battery. This in-

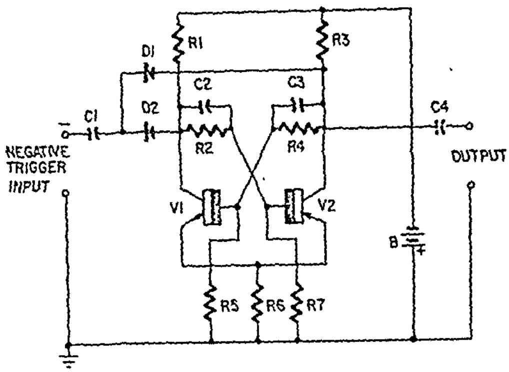


Fig. 708. Bistable circuit (flip-flop) with junction transistors.

creased voltage is coupled through R2–R7 to the base of V2, causing the latter's collector current to rise further. This action proceeds rapidly until V2 is conducting and V1 nonconducting. The next trigger pulse will flip conduction back to V1, and so on.

Since V2 is conducting only half of the time, an output pulse is delivered only in response to every second input-trigger pulse. This is true scale-of-two operation.

Recommended Reading

1. R. L. Trent, "Two Transistor Binary Counter," *Electronics*, July, 1952; p. 100.
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practical transistor circuits

THE circuits presented in this chapter show practical transistor applications. These circuits have been selected to illustrate their use in such common devices as amplifiers, oscillators, radio receivers, and measuring instruments. Each unit has been tested. The reader will think of other devices which can be made from combinations of the circuits given here.

Only a limited number of circuits can be presented in this single chapter. A few typical applications therefore have been chosen. For a more exhaustive supply of applications, see the author's book *Transistor Circuits*[1]. Additional practical material will be found in the articles mentioned in the reading list at the end of this chapter.

For best results, use the specified components and values. When these components are not readily available, substitutes should be used only when their electrical characteristics are identical to those specified.

Check all wiring and the polarities of transistors, diodes, electrolytic capacitors, dc meters, and batteries in the circuit before switching on the power to any device. This precaution will prevent damage to the components and will forestall faulty operation. The rules of good workmanship apply with particular emphasis to transistor equipment.

Single-stage R-C-coupled of amplifiers

Fig. 801 shows three single-stage audio amplifiers of the R-C-coupled type. Each is shown with a p-n-p transistor. N-p-n tran-

[1]Also published by Gernsback Library, Inc.

voltage lowers the base voltage of V1 (through R4–R5) and therefore the collector current of V1. This action causes the collector voltage of V1 to rise toward the potential of the battery. This in-

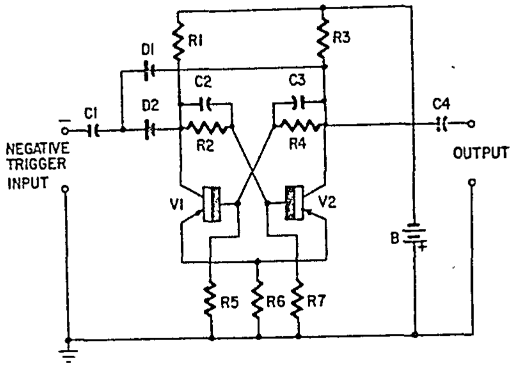

Fig. 708. *Bistable circuit (flip-flop) with junction transistors.*

creased voltage is coupled through R2–R7 to the base of V2, causing the latter's collector current to rise further. This action proceeds rapidly until V2 is conducting and V1 nonconducting. The next trigger pulse will flip conduction back to V1, and so on.

Since V2 is conducting only half of the time, an output pulse is delivered only in response to every second input-trigger pulse. This is true scale-of-two operation.

Recommended Reading

1. R. L. Trent, "Two Transistor Binary Counter," *Electronics,* July, 1952; p. 100.

2. B. G. Farley, "Dynamics of Transistor Negative-Resistance Circuits," *Proceedings of the IRE,* November, 1952; p. 1497.

3. A. W. Lo, "Transistor Trigger Circuits," *Proceedings of the IRE,* November, 1952; p. 1531.

4. E. Eberhard, R. O. Endres and R. P. Moore, "Counter Circuits Using Transistors," *RCA Review,* December, 1949; p. 459.

5. H. J. Reich and R. L. Ungvary, "Transistor Trigger Circuits," *Review of Scientific Instruments,* August, 1949; p. 586.

6. A. Eugene Anderson, "Transistors in Switching Circuits," *Proceedings of the IRE,* November, 1952; p. 1541.

7. L. P. Hunter and H. Fleisher, "Graphical Analysis of Some Transistor Switching Circuits," *Proceedings of the IRE,* November, 1952; p. 1559.

8. Robert L. Trent, "A Transistor Reversible Binary Counter," *Proceedings of the IRE,* November, 1952; p. 1562.

practical transistor circuits

THE circuits presented in this chapter show practical transistor applications These circuits have been selected to illustrate their use in such common devices as amplifiers, oscillators, radio receivers, and measuring instruments Each unit has been tested. The reader will think of other devices which can be made from combinations of the circuits given here.

Only a limited number of circuits can be presented in this single chapter. A few typical applications therefore have been chosen. For a more exhaustive supply of applications, see the author's book *Transistor Circuits*[1] Additional practical material will be found in the articles mentioned in the reading list at the end of this chapter.

For best results, use the specified components and values. When these components are not readily available, substitutes should be used only when their electrical characteristics are identical to those specified.

Check all wiring and the polarities of transistors, diodes, electrolytic capacitors, dc meters, and batteries in the circuit before switching on the power to any device This precaution will prevent damage to the components and will forestall faulty operation. The rules of good workmanship apply with particular emphasis to transistor equipment.

Single-stage R-C-coupled of amplifiers

Fig 801 shows three single-stage audio amplifiers of the R-C-coupled type. Each is shown with a p-n-p transistor N-p n tran-

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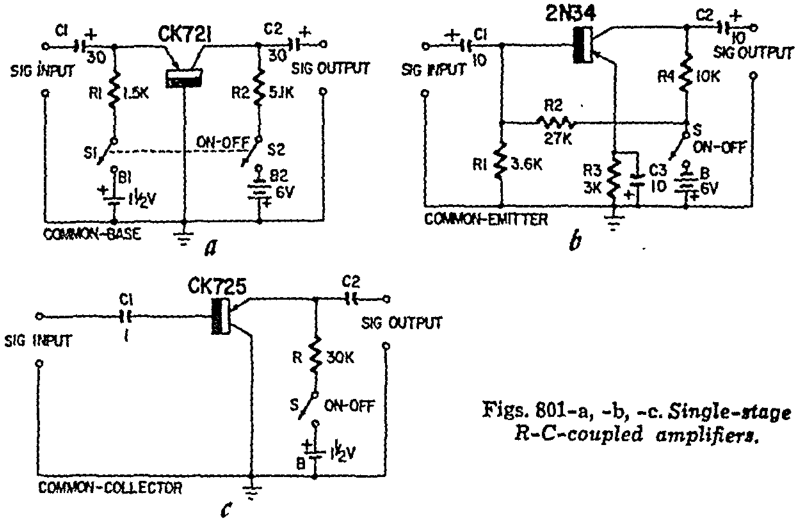
capacitors are reversed.

Fig. 801-a is a common-base circuit. When operated into a high-



Figs. 801-a, -b, -c. *Single-stage R-C-coupled amplifiers.*

impedance load (50,000 ohms or higher), this circuit provides a voltage gain of approximately 30. The input impedance is 130 ohms and the output impedance approximately 5,000 ohms. The maximum input-signal voltage which may be applied before output-signal peak clipping occurs is 0.1 volt rms. The corresponding maximum output-signal voltage is 3 volts rms. The frequency response is flat within —1 db from 50 cycles to 50 kc.

Fig. 801-b is a common-emitter circuit. It has an input impedance of 800 ohms, an output impedance of 10,000 ohms, and provides a voltage gain of 80 when operated into a high-impedance load. The maximum input-signal voltage is 20 mv rms and the corresponding maximum output-signal voltage is 1.7 volts rms, before output peak clipping occurs. The frequency response is flat within —2 db from 50 cycles to 10 kc.

Fig. 801-c is a common-collector circuit with an input impedance of 1 megohm at 1,000 cycles and an output impedance of 30,000 ohms. The voltage gain is 0.95 between 20 cycles and 10 kc and drops to 0.88 at 50 kc. The frequency response is flat within 0.7 db from 20 cycles to 50 kc.

Single-stage transformer-coupled af amplifiers

Fig. 802 shows three single-stage audio amplifiers of the trans-

former-coupled type. Transformer coupling provides higher power gain than is obtainable with R-C-coupled transistor amplifiers.

Fig. 802-a is a common-base circuit. This stage has an input

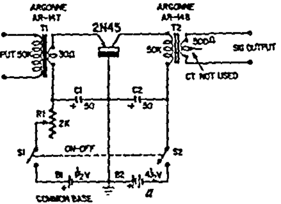


Fig. 802-a. *Transformer-coupled amplifier using common-base circuit.*

impedance of 50,000 ohms and output impedance of 500 ohms. The power gain is 27 db and power output is 2 mw. An input-signal driving power of 4 microwatts will give full output.

Fig. 802-b is a common-emitter circuit. Here, the power gain

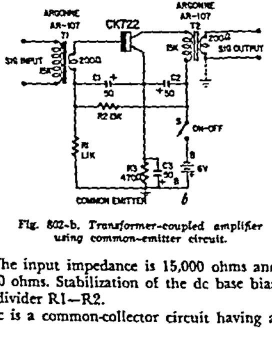


Fig. 802-b. *Transformer-coupled amplifier using common-emitter circuit.*

is 35 db. The input impedance is 15,000 ohms and output impedance 200 ohms. Stabilization of the dc base bias is provided by voltage divider R1—R2.

Fig. 802-c is a common-collector circuit having an input im-

pedance of 1 megohm and output impedance of 100 ohms. The power gain is approximately 12 db.

Multistage af amplifiers

Fig. 803 shows two multistage audio amplifiers. R-C coupling is employed throughout in Fig. 803-a while transformer coupling is used in Fig. 803-b. The higher power gain afforded by transformer coupling allows the use of fewer stages in the second circuit. Both circuits employ the common-emitter connection of transistors for highest gain.

The four-stage R-C-coupled circuit of Fig. 803-a provides an overall voltage gain of 4,000 when GAIN CONTROL R3 is set to maximum. The maximum input-signal voltage before output-signal peak clipping occurs is 0.2 mv rms. The corresponding

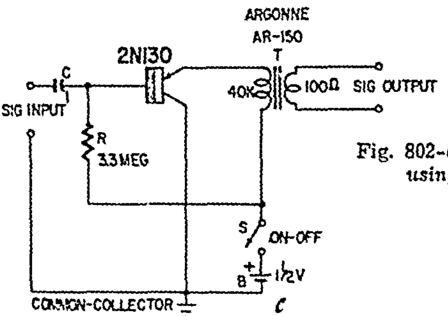


Fig. 802-c. *Transformer-coupled amplifier using common-collector circuit.*

maximum output-signal voltage is 1.2 rms. The input impedance of this amplifier is approximately 1,000 ohms and the output impedance 10,000 ohms. The frequency response is —8 db from 40 cycles to 10 kc. Current drain is approximately 10 ma dc.

The three-stage transformer-coupled circuit of Fig. 803-b has an input impedance of 1,000 ohms, an output impedance of 1,200 ohms and provides an overall power gain of 80 db when VOLUME CONTROL R5 is set to maximum. The power output is 6 mw. Current drain is 8 ma dc.

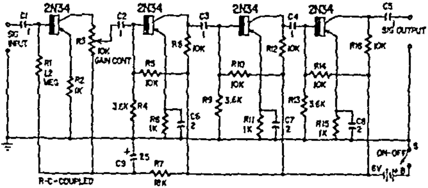
In each of the multistage amplifiers, a decoupling filter is provided to prevent motorboating. In Fig. 803-a this filter is composed of C9 and R7; in Fig. 803-b it is C3 and R4. In Fig. 803-a dc base-bias stabilization is provided by voltage dividers R4—R5, R9—R10, and R13—R14; and in Fig. 803-b by R1—R2, R6—R7, and R9—R10.

Microphone-case preamplifier

A single-stage R-C-coupled amplifier can be made small enough

to fit into the case of a small dynamic microphone. The circuit is seen in Fig. 804.

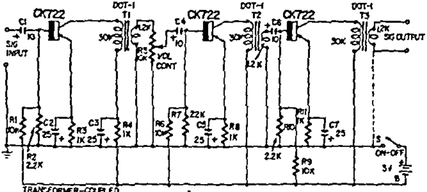
The voltage gain provided by this preamplifier allows a good-



a

Fig 803-a. R-C-coupled multistage audio amplifier.

quality dynamic microphone to be used in a mobile transmitter. Space and power requirements usually limit such transmitters



b

Fig 803-b Transformer-coupled multistage audio amplifier.

to poorer-grade carbon microphones, since the latter have higher voltage output and therefore require less speech amplification.

Loudspeaker-operating of power amplifiers

Fig. 805 gives the circuits of two push-pull audio power amplifiers suitable for driving loudspeakers. Both are class-B amplifiers. The circuit shown in Fig. 805-a delivers an output of 75 mw. This amplifier is adequate for portable radio receive

pedance of 1 megohm and output impedance of 100 ohms. The power gain is approximately 12 db.

Multistage af amplifiers

Fig. 803 shows two multistage audio amplifiers. R-C coupling is employed throughout in Fig. 803-a while transformer coupling is used in Fig. 803-b. The higher power gain afforded by transformer coupling allows the use of fewer stages in the second circuit. Both circuits employ the common-emitter connection of transistors for highest gain.

The four-stage R-C-coupled circuit of Fig. 803-a provides an overall voltage gain of 4,000 when GAIN CONTROL R3 is set to maximum. The maximum input-signal voltage before output-signal peak clipping occurs is 0.2 mv rms. The corresponding

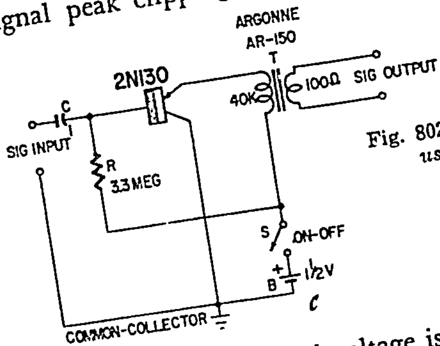


Fig. 802-c. *Transformer-coupled amplifie using common-collector circuit.*

maximum output-signal voltage is 1.2 rms. The input impeda of this amplifier is approximately 1,000 ohms and the out impedance 10,000 ohms. The frequency response is —8 db f 40 cycles to 10 kc. Current drain is approximately 10 ma

The three-stage transformer-coupled circuit of Fig. 803-b an input impedance of 1,000 ohms, an output impedance of ohms and provides an overall power gain of 80 db when vo CONTROL R5 is set to maximum. The power output is 6 Current drain is 8 ma dc.

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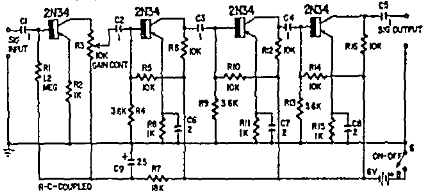
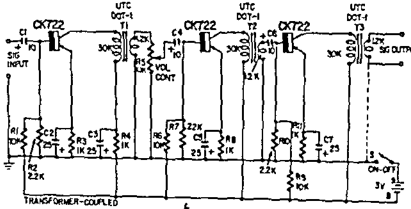
The voltage gain provided by this preamplifier allows a good-



a

Fig. 803-a *R-C-coupled multistage audio amplifier.*

quality dynamic microphone to be used in a mobile transmitter. Space and power requirements usually limit such transmitters



b

Fig. 803-b *Transformer-coupled multistage audio amplifier*

to poorer-grade carbon microphones, since the latter have higher voltage output and therefore require less speech amplification

Loudspeaker-operating of power amplifiers

Fig. 805 gives the circuits of two push-pull audio power amplifiers suitable for driving loudspeakers. Both are class B amplifiers. The circuit shown in Fig 805 a delivers an output of 175 mw. This amplifier is adequate for portable radio receivers and

intercoms. Power transistors are used in the circuit of Fig. 805-b which delivers 5 watts.

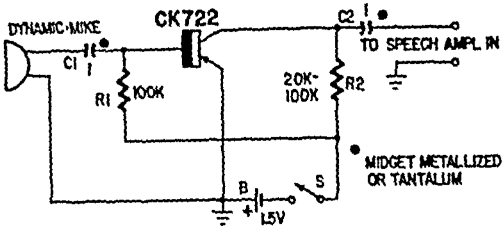


Fig. 804. *Subminiature amplifier for inclusion in mobile dynamic microphone.*

The low-level circuit of Fig. 805-a requires an input-signal driving power of 0.2 mw. The power gain is 30 db. The total zero-signal collector current is approximately 4 ma dc and the maximum-signal collector current is 26 ma.

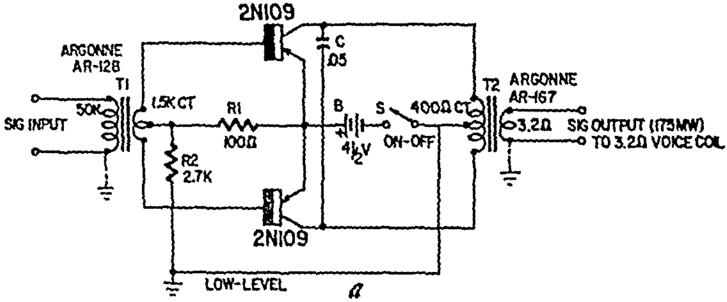


Fig. 805-a. *Low-level push-pull audio amplifier.*

The high-level circuit of Fig. 805-b requires an input signal driving power of 250 mw. Power gain is 20 db. The total zero-signal collector current is approximately 2 ma and the maximum-signal collector current 550 ma.

Hearing aid

Fig. 806 gives the circuit of a simple hearing aid that can be built from relatively inexpensive parts. This circuit has good sensitivity and output. The three-stage amplifier employs a low-noise transistor (type 2N133) in the input stage and high-alpha units (CK721) in the intermediate and output stages. Transformer coupling is used between stages for maximum overall power gain.

Switch S is integral with the subminiature ("dime-size") vol-

ume control potentiometer R4. The low-impedance magnetic microphone is connected directly into the base-emitter input circuit of the first transistor. Similarly, the magnetic earpiece is

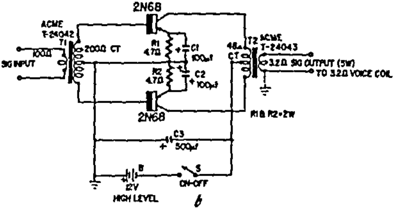


Fig. 805-b. *High-level push-pull audio amplifier.*

connected directly into the collector circuit of the output transistor.

The circuit is powered by a single cell. This component may

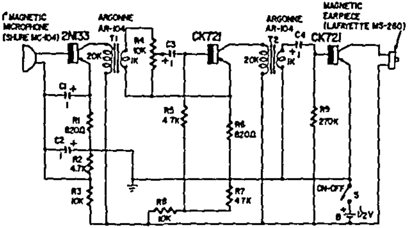


Fig. 806. *Simple hearing aid can be built economically*

be a 1.5-volt penlight cell or a small mercury cell. Good cell life is obtained.

Some adjustment of resistance R9 may be necessary for maximum output vs. minimum distortion.

intercoms. Power transistors are used in the circuit of Fig. 805-b which delivers 5 watts.

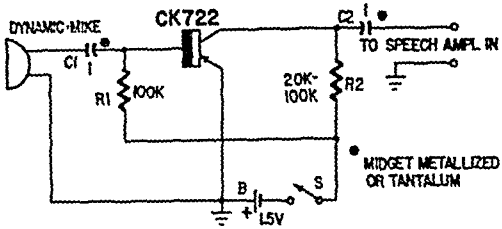


Fig. 804. Subminiature amplifier for inclusion in mobile dynamic microphone.

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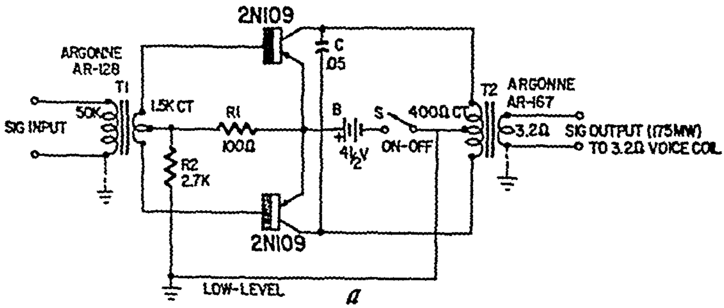

Fig. 805-a. Low-level push-pull audio amplifier.

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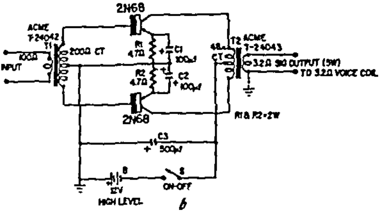

Fig. 805-b. High-level push-pull audio amplifier.

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The circuit is powered by a single cell. This component may

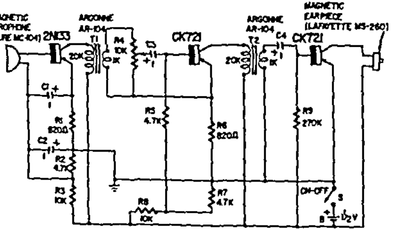

Fig. 806. Simple hearing aid can be built economically.

be a 1.5-volt penlight cell or a small mercury cell. Good cell life is obtained.

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Simplified single-stage hearing aid

For use by the not-too-hard of hearing, the very simple circuit of Fig. 807 offers interesting possibilities as an inexpensive hearing

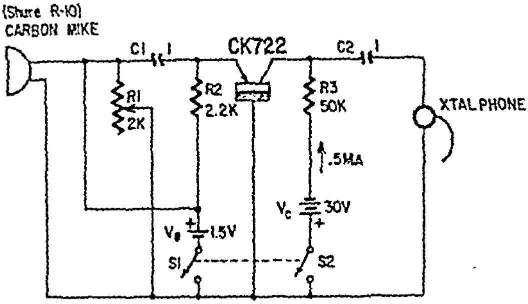

Fig. 807. *Single-stage experimental hearing aid.*

aid. This instrument may be employed to some extent also as a stethoscope, detectaphone and sound-pickup device where the high gain of the hearing aid shown in Fig. 806 is not required. Aside from lower gain, this circuit also has the disadvantage of

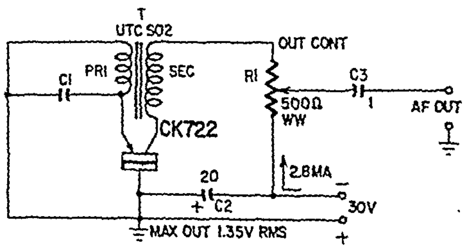

Fig. 808. *Audio oscillator–modulator.*

higher noise level as the result of hiss generated by the carbon microphone.

A single CK722 transistor is used in the grounded-base R-C coupled connection. The emitter bias battery, V_e, also supplies current to the carbon microphone. Microphone load resistor R1 is adjusted for best compromise between high gain and low noise level. The value of the collector load resistor R3 will vary somewhat with individual transistors and should be chosen for highest output signal.

Basic audio oscillator

Fig. 808 shows the circuit of a simple transformer-feedback audio oscillator using a CK722 transistor in the grounded-base

connection. A subminiature UTC type SO2 transformer (1-to-3 primary-to-secondary turns ratio) was used in the laboratory

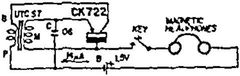

Fig. 809. Code-practice oscillator.

model, but any larger-sized unit with comparable characteristics can be used.

The operating frequency is determined chiefly by the values of capacitor C1 and the primary inductance of the transformer.

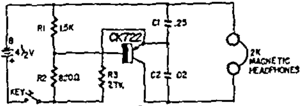

Fig. 810-a Colpitts oscillator using magnetic headphones as the tuning inductor

The transformer windings must be poled correctly for oscillation If oscillation is not obtained readily, reversing the connec-

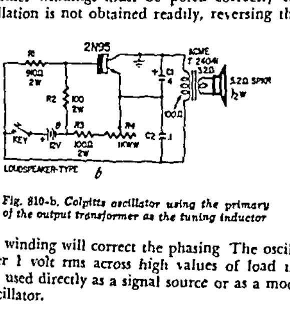

Fig. 810-b. Colpitts oscillator using the primary of the output transformer as the tuning inductor

tions of one winding will correct the phasing The oscillator will develop over 1 volt rms across high values of load impedance and may be used directly as a signal source or as a modulator in an rf test oscillator.

Low-drain code-practice oscillator

The ability of the junction transistor to operate with low values of dc input power is utilized in the code-practice oscillator shown in Fig. 809. A hearing-aid transformer will convert the instrument into a vest-pocket model. This oscillator operates on 14 microamperes supplied by a single 1.5-volt penlight cell. The current will vary one way or the other with individual transistors.

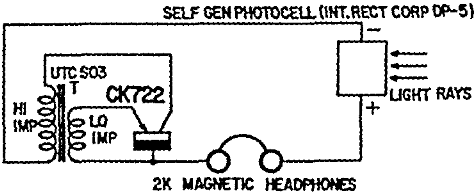

Fig. 811. *Light-powered audio oscillator.*

With the circuit constants shown, the signal frequency is approximately 700 cycles. The frequency may be increased by decreasing the value of C, and vice versa.

Colpitts type code-practice oscillators

The feedback transformer in the oscillator circuit of Fig. 809 is eliminated in the Colpitts circuits shown in Fig. 810.

In Fig. 810-a, the magnetic headphones supply the inductance required for the oscillator circuit. Capacitors C1 and C2 resonate this inductance to set the oscillator frequency. With a pair of 2,000-ohm headphones, the frequency was measured as 750 cycles with C1 equal to 0.25 μf and C2 to .02 μf. The frequency may be raised by decreasing C1 and C2 simultaneously, and lowered by increasing their values. At any chosen frequency, the ratio between the two capacitances must be approximately 10 to 1, as shown in Fig. 810-a.

The circuit given in Fig. 810-b provides loudspeaker operation at 0.5 watt output. This circuit is similar to the preceding one except that the inductance of the primary winding of the output transformer T acts with capacitances C1 and C2 to set the oscillator frequency. At the lowest-resistance setting of potentiometer R4, the frequency is 3,500 cycles. When R4 is set to 1,000 ohms, it is 360 cycles. Although the current drain from battery B is 170 ma when the key is closed, the code-practice service is so intermittent that eight 1.5-volt size-D flashlight cells in series will last quite long.

Light-powered audio oscillator

Another interesting oscillator application of the low dc requirement of the junction transistor is illustrated in Fig. 811. Here, a miniature transformer-feedback audio oscillator is operated on the direct current supplied by a self-generating photocell under illumination.

In subdued room light, a .02-millivolt rms audio signal was developed across 2,000-ohm magnetic headphones. A 100-watt lamp, 1 foot from the cell, gave 0.5 mv. Between 1 and 2 mv were obtained in direct sunlight. The frequency was 900 cycles, but can be lowered by connecting a suitable capacitance in parallel with either the primary or secondary of the transformer. The transformer windings must be poled correctly for oscillation. In lieu of headphones, the oscillator output, developed across a 2,000-ohm resistor, may be presented to an audio amplifier for measure-

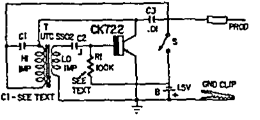

Fig 812. *Audio signal injector for amplifier servicing.*

ment or control purposes. Low-power-drain oscillators of this type have been operated also from the direct current obtained from thermocouples, charged capacitors and similar sources.

Audio signal injector

A *signal injector* is convenient for introducing a test signal at various points in an audio amplifier during troubleshooting. A battery-operated pocket-sized instrument gives the convenience of portability when a full-sized oscillator could not be handled at the test location. A transistor type injector allows the compactness and economy of a single-cell power supply.

The signal-injector circuit shown in Fig. 812 employs a CK722 transistor as a grounded-emitter transformer-feedback oscillator. The transformer is a UTC type SSO2 subsubminiature hearing-aid unit. A single 1.5-volt penlight cell supplies power. The entire instrument can be assembled into the shell of a test probe.

The transformer windings must be poled correctly for oscilla-

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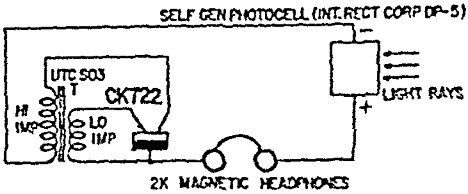

Fig. 811. *Light-powered audio oscillator.*

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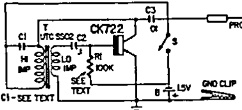

Fig 812. *Audio signal injector for amplifier servicing.*

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dio signal injector

A *signal injector* is convenient for introducing a test signal at rious points in an audio amplifier during troubleshooting. A ttery-operated pocket-sized instrument gives the convenience portability when a full-sized oscillator could not be handled the test location. A transistor type injector allows the compact-ss and economy of a single-cell power supply.

The signal-injector circuit shown in Fig. 812 employs a CK722 nsistor as a grounded-emitter transformer-feedback oscillator. he transformer is a UTC type SSO2 subsubminiature hearing-d unit. A single 1.5-volt penlight cell supplies power. The tire instrument can be assembled into the shell of a test probe. The transformer windings must be poled correctly for oscilla-

tion. The 100,000-ohm resistor will be correct in most cases but may be increased or decreased to match individual transistors. It should be adjusted for a collector current lower than 0.5 ma dc. The signal frequency can be adjusted in steps by fixed capaci-

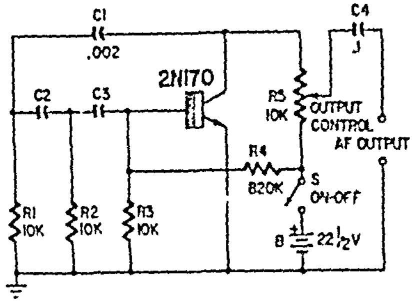

Fig. 813. *Phase-shift audio oscillator.*

tors in parallel with either the primary or secondary of the transformer.

Phase-shift af oscillator

The phase-shift oscillator circuit has the advantage that it requires neither transformers nor coils. The frequency is determined solely by the constants of a resistance–capacitance network. This type of oscillator is well known in vacuum-tube practice.

Fig. 813 shows the circuit of a 2,000-cycle phase-shift oscillator employing a single n-p-n transistor. The frequency is set by the three-leg phase-shift network C1R1–C2R2–C3R3. The total harmonic distortion is 0.26%. The frequency may be changed by substituting other R and C values in the phase-shift network. At any chosen frequency, the three resistances (R1, R2, R3) must be equal as well as the three capacitances (C1, C2, C3).

Potentiometer R5 gives smooth control of the af output. The circuit is designed to operate into a high impedance, such as a vacuum-tube amplifier or oscilloscope.

Total current drain from the 22.5-volt battery B is approximately 1.1 ma dc.

100-kc crystal oscillator

Fig. 814 shows a 100-kc crystal oscillator circuit suitable for use as a secondary frequency standard. The tuned circuit is resonated to the crystal frequency by adjustment of the slug screw of the 0.5–5-mh inductor L. This circuit is very active and can be keyed.

The LC2 tank may be tuned over a frequency range of approximately 70 to 220 kc. Any crystal within this range will be accommodated.

High-frequency crystal oscillator

The low-drain circuit shown in Fig. 815 will oscillate at crystal

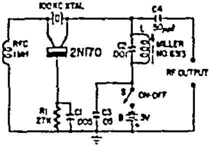

Fig 814. 100-kc crystal oscillator.

frequencies up to 7 mc. Higher-frequency operation may be obtained with a transistor type having a higher alpha-cutoff frequency that that of the 2N94A shown

The inductance of coil L is chosen to resonate with tuning

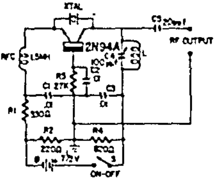

Fig 815 High-frequency crystal oscillator

capacitor C4 at the crystal frequency. Coil-winding data may be found in amateur handbooks. The rf output voltage into a high-impedance load is approximately 5 volts rms.

While high-impedance capacitance-coupled output (through C5) is shown, low-impedance output may be obtained by a link-coupled coil of a few turns closely coupled to the ground end of L.

All-band rf oscillator

A drift transistor (type 2N247) is used in the self-excited rf oscillator circuit shown in Fig. 816. This circuit has been adapted from an original RCA layout. By proper choice of the inductance of plug-in coils (L), this circuit may be tuned as high as 50 mc.

High-impedance capacitance-coupled output (through C5) is

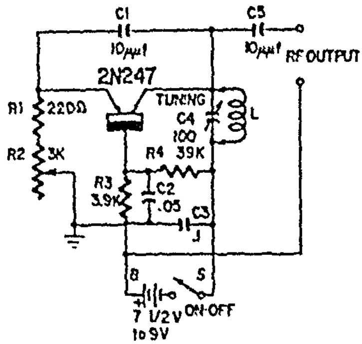

Fig. 816. *All-wave rf oscillator.*

shown. However, low-impedance output has less reaction on the oscillation frequency and may be obtained with a small link coil closely coupled to the ground end of coil L.

The rf output tends to fall off as tuning capacitor C4 is varied. This effect may be counteracted by adjustment of potentiometer R2.

Diode broadcast receiver with transistor amplifier

Fig. 817 shows the circuit of a simple diode (crystal) broadcast tuner followed by a two-stage junction transistor amplifier. The single power supply for this receiver is a 1.5-or 3-volt battery. The audio amplifier is transformer-coupled for best interstage impedance match and for highest audio gain. Grounded-emitter stages are used. The 100,000-ohm base resistors R1 and R2 must be adjusted for individual transistors. The proper value will limit the no-signal collector current to approximately 100 micro-amperes dc.

The input coupler (L1–L2) is a standard broadcast type antenna input coil, such as Miller type 20-A. Tuning is accomplished entirely by means of the 365-μμf variable capacitor. The audio output into 2,000-ohm magnetic headphones is approximately 2 mw, depending upon the strength of the received radio signal. This output may be increased by boosting the battery voltage, provided the maximum collector dissipation of 30 mw is not

exceeded. The loudspeaker-operating audio output amplifier stage (Fig. 805-a) may be added in place of the headphones. This receiver requires a good outside antenna and a solid ground connection for best performance.

Additional stages of audio amplification will greatly boost the

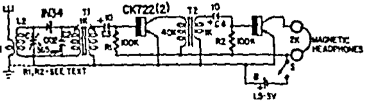

Fig. 817. *Diode type broadcast receiver with transistor af amplifier.*

sensitivity of the receiver. Since this circuit has no rf amplification, gain is supplied by the audio amplifier alone. If a multistage af amplifier such as the circuit shown in Fig. 803 b, is employed

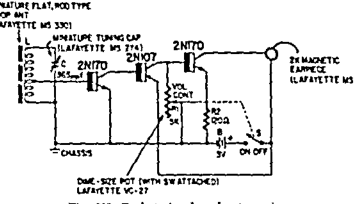

Fig 818 Pocket-size broadcast receiver.

and this is followed by a loudspeaker-operated stage, such as Fig. 805 a, the circuit will have considerable sensitivity. The L1-L2 coupler then may be replaced with a ferrite loop type antenna and most local broadcast stations received without an outside antenna or ground.

Pocket broadcast receiver

The circuit of a simple broadcast receiver which can be built small enough to be carried in a vest pocket is shown in Fig 818 This circuit, developed by General Electric Co, employs a transistor detector direct-coupled to a two-stage direct-coupled audio amplifier. By using an n-p-n transistor in the detector stage and

p-n-p's in the amplifier, a single battery is permissible although the circuit is direct-coupled. A pair of 1.5-volt penlight cells connected in series will serve as battery B.

While this receiver does not have the sensitivity of a transistor superheterodyne, it will give good performance on strong local stations. For increased pickup, an outside antenna may be connected to the top of the loop antenna coil L.

Phone monitor

The simple diode detector with headphones has been used as a monitor of radio signals since the inception of the germanium diode. With this inexpensive device, aural monitoring of signal quality is accomplished easily. Diode monitors are found in commercial as well as amateur stations.

A disadvantage of the diode type monitor has been its low audio output except when operated dangerously close to the radio transmitter. While small-sized audio amplifiers have been added on occasion, there is general objection to these additions on the grounds that ac amplifiers destroy the desirable isolation feature of the instrument and battery-operated amplifiers have not been economical.

The low power requirements of the junction transistor adapt it to use as a single-stage audio headphone amplifier in the diode type monitor. A circuit is shown in Fig. 819. Here, a grounded-emitter CK722 is transformer-coupled to the monitor diode output. Coil L and capacitor C1 are chosen to tune to the carrier

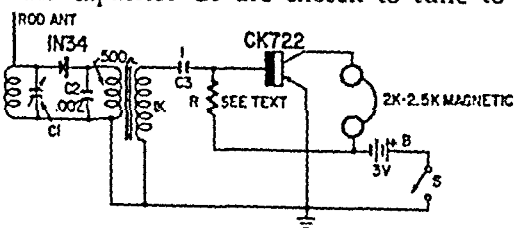

Fig. 819. Monitor for radiophone transmission.

frequency of the transmitter. A short rod antenna or length of insulated wire usually will provide enough pickup anywhere in and around the transmitter building. Resistor R will have a value of approximately 100,000 ohms but must be adjusted for a no-signal collector current of not more than 100 microamperes.

Sensitive field-strength meter

Another popular use of the diode detector at radio transmitting

stations has been as a simple field-strength meter. This instrument, built around a pickup antenna, germanium diode and dc milliammeter, is invaluable for carrier monitoring and for making comparative intensity measurements in transmitter and antenna adjustments. The sensitivity of the diode type field-strength meter is increased by the use of a low-range dc microammeter in place of the usual milliammeter. But the microammeter is expensive and often less rugged mechanically.

Fig. 820 shows how a grounded-emitter junction transistor may be used as a dc amplifier between the diode output and a 0–1 dc milliammeter. A CK722 transistor will provide a base-to-collector

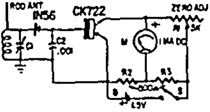

Fig. 820 Sensitive field-strength meter.

current amplification (*beta*) of 10 to 12; a CK721, 30 to 40. Thus, the amplifier circuit converts the 0–1 milliammeter into a 0–100 dc microammeter with a CK722, and into a 0–25 microammeter with a CK721.

The dc output of the 1N56 diode rectifier passes through the base-emitter input path of the transistor. The diode is poled so that negative voltage is applied to the base of the transistor. Capacitor C2 is for radio-frequency bypassing. The values of coil L and variable capacitor C1 are chosen for resonance at the transmitter carrier frequency. With no rf signal input, the meter is set initially to zero by adjusting the zero-set rheostat R1.

Amplifier type dc microammeter

The dc amplifier and indicating meter circuit of the field-strength meter has been separated in Fig 821 for use as a sensitive dc microammeter. Using a single CK721 or CK722 grounded emitter transistor, the 0–1 dc milliammeter will read 0–25 dc microamperes with a CK721, or 0–100 microamperes with a CK722. A 0–100 dc microammeter at M will read 0–2.5 μa with a CK721, or 0–10 μa with a CK722.

With no input signal voltage, the meter is set initially to zero, as in a vacuum-tube voltmeter, by adjustment of the zero-set rheostat R2. The dc input terminals can be open or shorted for this adjustment. After zero-setting, the circuit is standardized

initially by applying an accurately known dc input current corresponding to full-scale deflection of the meter and by adjusting the calibration control rheostat R1 for full-scale deflection. When a 0–1 milliammeter is used, this input current should be 25 μa

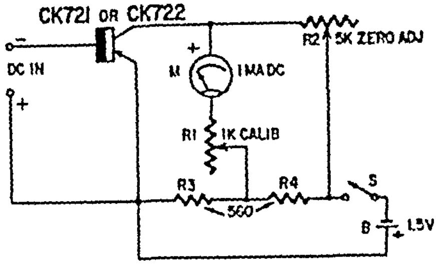

Fig. 821. *Amplifier type dc microammeter.*

for the CK721 or 100 μa for the CK722. With a 0–100 microammeter, a 2.5-μa calibration current is used with CK721 and 10 μa with the CK722.

Linearity of the instrument is very good. However, an individual calibration should be made when best accuracy is desired. Total drain from the single 1.5-volt cell is 1.5 ma at full-scale deflection of the 0–1 milliammeter.

Sensitive dc relay

The grounded-emitter junction-transistor dc amplifier is applicable also for building up small direct currents to levels sufficient to operate a milliampere type dc relay. Thus microampere response may be obtained with a more rugged milliampere relay. Fig. 822 shows the simple circuit which can be used for this purpose.

Best results are obtained with relays having low coil resistance. The collector voltage is chosen such that the steady collector

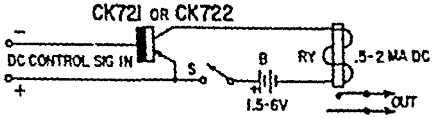

Fig. 822. *Amplifier type sensitive dc relay.*

current flowing through the relay coil is somewhat less than required to pick up the relay. A few microamperes applied to the input terminals of the circuit then will actuate the relay. Highest sensitivity is afforded by high-alpha junction transistors, such as

the CK721. The latter provides a base-to-collector amplification of 30 to 40. *Current amplification with the CK722 can reach a maximum* of 12.

Heterodyne frequency meter

The tube type heterodyne frequency meter is a familiar instrument in laboratories and radio stations for the measurement of radio frequencies by *the zero-beat method. The conventional*

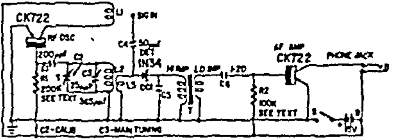

Fig. 823. Transistor heterodyne frequency meter.

instrument consists of (1) a single-band, unmodulated, tuned rf oscillator; (2) aperiodic detector *or* mixer and (3) audio "beat-note" amplifier.

The signal of unknown frequency is applied to the detector. The oscillator frequency then is adjusted to zero-beat with the signal or one of its harmonics. An oscillator harmonic also may zero-beat the signal. At zero beat, the frequency is read from the oscillator dial, and the unknown determined *from this reading* and the proper harmonic number.

The advantages of a transistor type heterodyne frequency meter are (1) *complete isolation,* (2) *small size* and portability, (3) economy of operation not possible with battery type vacuum tubes and (4) absence of heating.

Fig. 823 shows the circuit of a transistor heterodyne frequency meter. The local oscillator employs a CK722 transistor in a grounded-emitter tickler feedback circuit. This oscillator tunes from 500 to 1,000 kc and allows identification of signals from 50 kc to 30 mc. Coil L2 consists of 115 turns of No. 32 enameled wire close-wound on a 1-inch diameter form. L1 *is 40 turns of No 26* enameled wire closewound on top of L2 and insulated from the latter. Output coupling coil L3 consists of 15 turns of insulated hookup wire jumble-wound *and cemented inside of the form* on which L2 is wound. Resistor R1 may need adjustment to an individual transistor to insure continuous oscillation throughout

the 500–1,000-kc band. The 365-μμf variable capacitor C3 is the main tuning control. Capacitor C2 is a calibration trimmer, which is set for zero beat when C3 is tuned to a harmonic of a 100-kc frequency standard. The 100-kc standard is coupled into the detector by direct connection to the signal input terminal.

The detector–mixer is a 1N34 diode. The test signal is applied through capacitor C4 to the diode simultaneously with the local oscillator signal. The detector is transformer-coupled to a grounded-emitter audio amplifier employing a CK722 transistor. In this latter circuit, the base resistor R2 will be approximately 100,000 ohms but must be adjusted with an individual transistor to limit the no-signal collector current to a maximum of 100 microamperes dc. Total current drain from a 15-volt hearing-aid-type battery is less than 1 ma dc.

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tests and measurements

THE measurement of dc voltages and currents at emitter, collector, and base electrodes is fundamental in determining transistor operating characteristics. Equipment required for these tests are a high-resistance dc vacuum-tube voltmeter (minimum input resistance of 10 megohms), low-resistance current meter (dc milliammeter or microammeter, as current level demands) and separate, adjustable constant-current dc supplies. While it is desirable in some tests to have separate voltmeters and current meters for each electrode, economy often dictates that switching arrangements be worked out for transferring a single pair of instruments between parts of the test circuit.

In testing transistors, a given current is passed into the electrode of interest. The resulting voltage then is measured between that electrode and the *reference electrode* (usually base or emitter). This test is always made at some specified constant value of current in the third electrode. For example, a current of 2 ma dc from one bias supply is passed through the emitter-to-base circuit of a transistor, while holding constant at 10 ma dc the current (furnished by a second supply) through the collector-to-base circuit. The corresponding collector-to-base voltage then is measured.

The collector polarity is negative and emitter positive in point-contact transistors made with n-type germanium, and p-n-p junction transistors. The collector polarity is positive and [illegible] negative in n-p-n junction units. Throughout this c[illegible] larities shown in the drawings are correct for n-type po[illegible] and p-n-p junction units. When checking n-p-n ju[illegible] type point-contact units, reverse all bias polarities [illegible]

Transistors normally are tested in the grounded-base or grounded-emitter circuit. In the grounded-base circuit, the emitter is the input electrode, the collector the output electrode, and emitter and collector voltages are referred to the base. In the grounded-emitter circuit, the base is the input electrode, and base and collector voltages are referred to the emitter. In the grounded-base test circuit, emitter and collector currents are measured, while in the grounded-emitter circuit, base and collector currents are measured.

It is insufficient to specify *only* current and voltage for a transistor electrode. For the information to be complete, one must state also at what value of current in the opposite electrode the measurements apply. This must be done even when the reference current is zero. Thus, it is meaningless to say only that collector current is 1.5 ma and collector-to-base voltage is 10 in a grounded-base test circuit. We must indicate that these values are obtained when the emitter current is 0.1 ma.

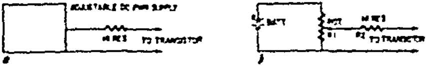
Constant-current dc supplies

The constant-current type of dc bias supply is essential in transistor testing, especially for furnishing bias to the emitter or base input electrode. If a specially designed supply unit of this type is not available, constant current may be obtained from a dc voltage supply or battery in the manner shown in Fig. 901. In each example, the voltage of the source is considerably higher than that required at the transistor electrode, and a high resistance is connected in series with one of the output terminals. This resistance, instead of the transistor parameters, consequently determines the magnitude of current.

In Fig. 901-a, an adjustable-output dc power supply is used. The output current is varied in magnitude by adjusting the output voltage of the supply. In Fig. 901-b a battery and potentiometer (R1) are used as the adjustable voltage source, and a high resistance (R2) is connected in series with one output lead for current regulation. The series resistance in each circuit should be as high as practicable in value (100 to 1,000 times the transistor resistance into which the current is introduced). This means that the source voltage also must be high. As an example, a 100-volt supply in series with 25,000 ohms can be used to supply constant emitter currents up to 2 ma. The potential at the emitter would be of the order of 0.2 volt.

Regarding constant-current supplies, a word of caution is in order. When the transistor is disconnected, the entire supply volt-

age appears at the current terminals. This can constitute a shock hazard, especially when supply voltages of several hundred volts



Figs. 901-a, -b. Adjustable dc power supply with current-regulating resistor (a). Battery with potentiometer and current-regulating resistor (b).

are employed. The operator must be careful that the power is switched off before removing the transistor or current leads from the test circuit.

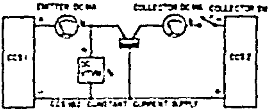


Fig. 902-a. Test circuit for measurement of input characteristics

Two constant-current dc bias supplies are required in transistor testing—one for the collector and the other for the emitter or base.

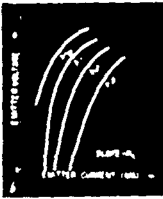


Fig. 902-b. Input characteristic curves

Measurements for characteristic curves

Figs. 902 to 905 show test-circuit setups for taking current

Transistors normally are tested in the grounded-base or grounded-emitter circuit. In the grounded-base circuit, the emitter is the input electrode, the collector the output electrode, and emitter and collector voltages are referred to the base. In the grounded-emitter circuit, the base is the input electrode, and base and collector voltages are referred to the emitter. In the grounded-base test circuit, emitter and collector currents are measured, while in the grounded-emitter circuit, base and collector currents are measured.

It is insufficient to specify *only* current and voltage for a transistor electrode. For the information to be complete, one must state also at what value of current in the opposite electrode the measurements apply. This must be done even when the reference current is zero. Thus, it is meaningless to say only that collector current is 1.5 ma and collector-to-base voltage is 10 in a grounded-base test circuit. We must indicate that these values are obtained when the emitter current is 0.1 ma.

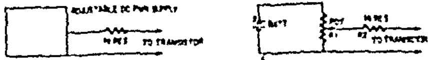
Constant-current dc supplies

The constant-current type of dc bias supply is essential in transistor testing, especially for furnishing bias to the emitter or base input electrode. If a specially designed supply unit of this type is not available, constant current may be obtained from a dc voltage supply or battery in the manner shown in Fig. 901. In each example, the voltage of the source is considerably higher than that required at the transistor electrode, and a high resistance is connected in series with one of the output terminals. This resistance, instead of the transistor parameters, consequently determines the magnitude of current.

In Fig. 901-a, an adjustable-output dc power supply is used. The output current is varied in magnitude by adjusting the output voltage of the supply. In Fig. 901-b a battery and potentiometer (R1) are used as the adjustable voltage source, and a high resistance (R2) is connected in series with one output lead for current regulation. The series resistance in each circuit should be as high as practicable in value (100 to 1,000 times the transistor resistance into which the current is introduced). This means that the source voltage also must be high. As an example, a 100-volt supply in series with 25,000 ohms can be used to supply constant emitter currents up to 2 ma. The potential at the emitter would be of the order of 0.2 volt.

Regarding constant-current supplies, a word of caution is in order. When the transistor is disconnected, the entire supply volt-

age appears at the current terminals. This can constitute a shock hazard, especially when supply voltages of several hundred volts



Figs. 901-a, -b. *Adjustable dc power supply with current-regulating resistor (a). Battery with potentiometer and current-regulating resistor (b).*

are employed. The operator must be careful that the power is switched off before removing the transistor or current leads from the test circuit.

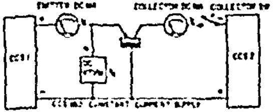


Fig. 902-a. *Test circuit for measurement of input characteristics*

Two constant-current dc bias supplies are required in transistor testing—one for the collector and the other for the emitter or base.

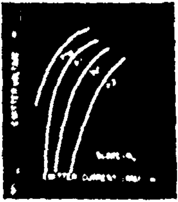


Fig. 902-b. *Input characteristic curves*

Measurements for characteristic curves

Figs. 902 to 905 show test-circuit setups for taking current and

voltage points for transistor characteristic curves. The test procedures are described separately in the following paragraphs.

Input characteristic

This is a plot of emitter current vs. emitter voltage for a constant value of collector current. The test circuit is shown in Fig. 902-a.

Select several values of collector current (i_c) at which measurements will be made. Take points for the first curve at zero col-

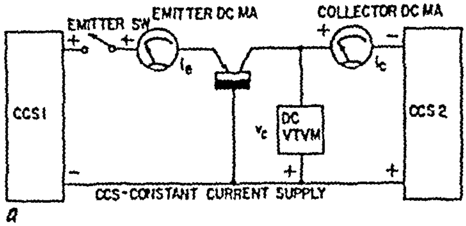


Fig. 903-a. *Test circuit for measurement of output characteristics.*

lector current by opening collector switch SW temporarily. Vary the emitter current (i_e) between the desired steps by adjusting the

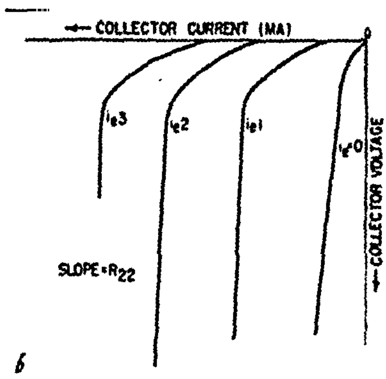


Fig. 903-b. *Output characteristic curves.*

output of constant-current supply 1, and record the corresponding emitter voltage (v_e) points. A curve similar to the top one (i_c = zero) in Fig. 902-b will be obtained. Close the switch and set the collector current to value i_c1 by adjusting the output of constant-current supply 2. Vary the emitter current in steps, as before, and

record corresponding emitter voltage points while holding the collector current constant. Repeat this procedure for constant collector-current values i,2, i,3, etc. A family of curves similar to Fig 902 b will be obtained.

Output characteristic

This is a plot of collector current vs. collector voltage for a

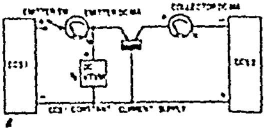


Fig 904-a. Test circuit for measurement of feedback characteristics

constant value of emitter current. The test circuit is shown in Fig 903 a.

Select several values of emitter current (i_e) for the measure-

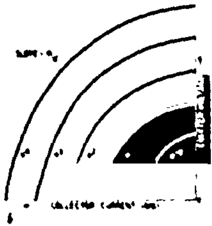


Fig 904-b. Feedback characteristic curves

ments. Take points for the first curve at zero emitter current by opening the emitter switch S temporarily, varying the collector current (i_c) by adjustment of constant-current supply 2, and recording corresponding collector voltage (v_c) points. A curve

similar to the lowest one ($i_e = 0$) in Fig. 903-b will be obtained. Close the switch and set the emitter current to value i_e1 by adjusting the output of constant-current supply 1. Vary the

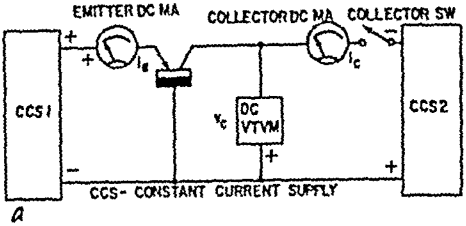


Fig. 905-a. *Test circuit for measurement of forward characteristics.*

collector current in steps, as before, and record corresponding collector voltage points while holding the emitter current con-

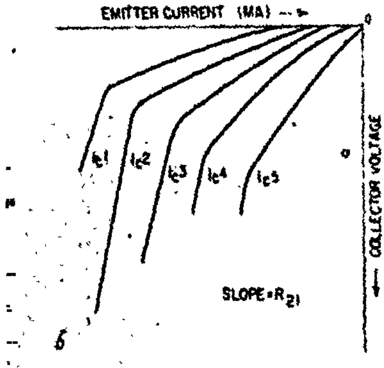


Fig. 905-b. *Forward characteristic curves.*

stant. Repeat this procedure for constant emitter-current values i_e2, i_e3, etc. A family of curves similar to Fig. 903-b will be obtained.

Feedback characteristic

This is a plot of collector current vs. emitter voltage for a constant value of emitter current. The test circuit is shown in Fig. 904-a.

Select several values of emitter current (i_e) at which measurements will be made. Take points for the first curve at zero emitter current by opening the emitter switch SW temporarily, varying

the collector current (i_c) by adjustment of constant-current supply 2, and recording the corresponding emitter voltage (v_e) points. A curve similar to the lowest one in Fig. 904-b ($i_e = 0$) will be obtained. Close the switch and set the emitter current to value

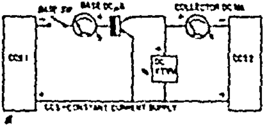


Fig. 904-a. Test circuit for measurement of grounded-emitter output characteristic.

i_b1 by adjusting the output of constant-current supply 1. Vary the collector current in steps, as before, and record corresponding emitter voltage points while holding the emitter current constant.

Fig. 904-b. Output characteristic curves of grounded emitter

stant. Repeat this procedure for constant emitter-current values of i_b2, i_b3, etc. A family of curves similar to Fig. 904-b will be obtained.

Forward characteristic

This is a plot of emitter current vs. collector voltage for a constant value of collector current. The test circuit is shown in Fig. 905-a.

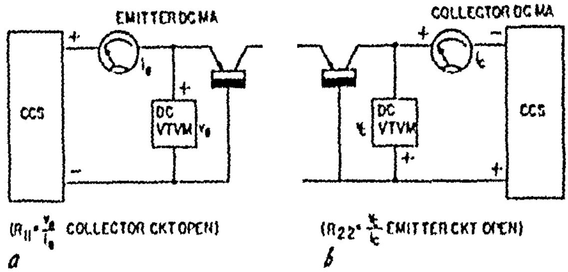
Select several values of collector current (i_c) at which measure-

ments will be made using constant-current supply 2. Take points for the first curve at the constant collector-current value of i_c1, varying the emitter current (i_e) by adjustment of the output of constant-current supply 1, and recording the corresponding collector voltage (v_e) points while holding the collector current constant. A curve similar to the top one in Fig. 905-b (i_c1) will be obtained. Repeat this procedure for constant collector-current values i_c2, i_c3, etc. A family of curves similar to Fig. 905-b will be obtained.

Output characteristic of grounded-emitter

This is a plot of collector current vs. collector voltage for a constant value of base current. The test circuit is shown in Fig. 906-a.

Select several values of base current (i_b) at which measurements will be made. Take points for the first curve at zero base



Figs. 907-a, -b. Input resistance measurement (a). Output resistance measurement (b).

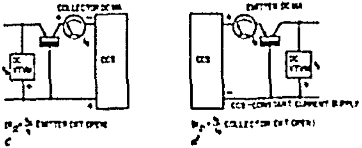
current by opening the base switch SW temporarily, varying the collector current (i_c) in steps by adjustment of constant-current supply 2, and recording the corresponding collector voltage (v_c) points. A curve similar to the lowest one in Fig. 906-b ($i_b = 0$) will be obtained. Close the switch and set the base curent to value i_b1 by adjustment of constant-current supply 1. Vary the collector current in steps, as before, and record the corresponding collector voltage points while holding the base current constant. Repeat this procedure for constant base-current values i_b2, i_b3, etc. A family of curves similar to Fig. 906-b will be obtained.

Input, feedback and forward characteristics also can be checked for the grounded-emitter junction transistor. Use the procedures outlined earlier, except substitute the base for the emitter in each instance, and observe the current-supply polarities shown in Fig. 906-a.

Determination of transistor resistances

The input, output, and transfer resistances of a transistor may be determined from dc characteristic curves plotted in the manner described in the preceding section.

The values of *input resistance* R_{11} may be found by measuring the slopes of the curves in Fig. 902 b, the *output resistance* R_{22}



Figs. 907-c, -d. Feedback resistance measurement (c) Forward resistance measurement (d)

from the slopes of the curves in Figs. 905 b and 906 b, *feedback resistance (base resistance)* R_{12} from the slopes of the curves in Fig. 904 b and forward transfer resistance R_{21} from the slopes of the curves in Fig. 905 b From these resistance values, emitter resistance (r_e), collector resistance (r_c), mutual resistance (r_m) and current amplification factor (alpha) can be calculated. For example:

$$r_e = R_{11} - R_{12}$$
$$r_c = R_{22} - R_{12}$$
$$r_m = R_{21} - R_{12}$$
$$\text{Alpha} = R_{21}/R_{22}.$$

The slopes of the curves of the dc characteristic give dynamic values of the transistor resistances and are to be preferred. However, the resistances also can be checked statically by means of single point current and voltage measurements with zero current in the reference electrode (input or output circuit open) For some purposes, resistance values obtained by this method will be sufficient.

The test circuits for static measurements are shown in Fig 907 The formula for the computation of the applicable resistance appears below each circuit.

It should be explained that the high input resistance of the dc vacuum tube voltmeter allows the emitter circuit in Fig. 907-c and the collector circuit in Fig. 907-d to appear open. How-

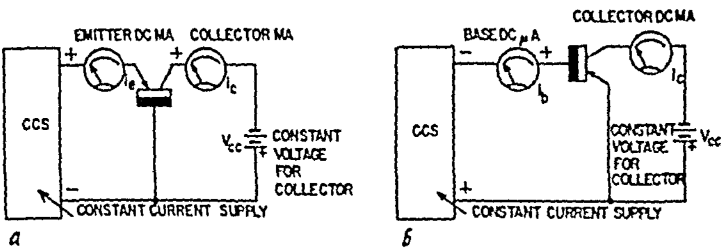
ever, the high collector-to-base terminal resistance of some junction transistors may be significant with respect to the voltmeter resistance and might limit the accuracy of measurement in Fig. 907-d.

Direct measurement of gain

Current amplification in transistors can be checked by comparing the output-current change with the small input-current change which produces it. The transistor output electrode voltage is held constant. Both the emitter-to-collector amplification (alpha) and the base-to-collector amplification (beta) can be checked in this manner.

Alpha measurement

The test circuit is shown in Fig. 908-a. The collector supply voltage (V_{cc}) and emitter current (i_e) levels at which the test will be made are chosen from the transistor characteristics sheet. In the test, i_e will be varied a small amount above and below the selected operating value, and the corresponding i_c change noted.



Figs. 908-a, -b. *Test circuit for current amplification, alpha (a). Test circuit for current amplification, beta (b).*

By adjustment of the output of the constant-current supply, set i_e to a value 0.1 ma lower than the selected emitter-current value. Record this setting as i_e1 and the corresponding collector current as i_c1. Next, set the emitter current to a point 0.1 ma higher than the selected value i_e. Record this new value as i_e2 and the corresponding collector current as i_c2. Alpha then is determined from the ratio of the collector and emitter current changes. Thus: $\alpha = (i_c2 - i_c1) / (i_e2 - i_e1)$.

Alpha measurements may be made at a variety of i_e and V_{cc} values to obtain complete data showing the variation of current amplification with these parameters.

Beta measurement

The test circuit for this measurement is shown in Fig. 908-b.

The collector supply voltage (V_{ce}) and base current (i_b) levels at which the test will be made are chosen from the transistor characteristics sheet. In the test, i_b will be varied a small amount above and below the selected value, and the corresponding change in i_c noted.

By adjustment of the output of the constant-current supply, set i_b to a value 20 microamperes below the selected i_b value. Record this setting as i_b1, and the corresponding collector current as i_c1.

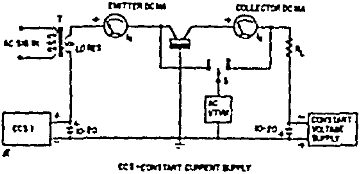


Fig. 909-a. Setup for testing the voltage amplification of the grounded-base circuit

Next, set the base current to a point 20 μa higher than the selected value, i_b. Record this new setting as i_b2, and the corresponding collector current as i_c2. Beta then is determined from the ratio

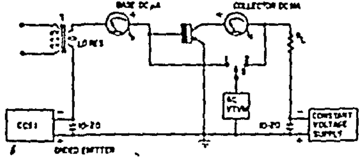


Fig. 909-b. Setup for testing the voltage amplification of the grounded-emitter circuit.

of the collector and base current changes. Thus: $\beta = (i_c2 - i_c1) / (i_b2 - i_b1)$. Beta measurements can be made at a variety of i_b and V_{ce} values to obtain complete data showing the variation of current amplification with the parameters.

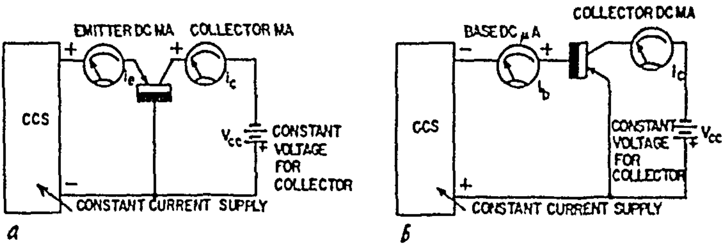
ever, the high collector-to-base terminal resistance of some junction transistors may be significant with respect to the voltmeter resistance and might limit the accuracy of measurement in Fig. 907-d.

Direct measurement of gain

Current amplification in transistors can be checked by comparing the output-current change with the small input-current change which produces it. The transistor output electrode voltage is held constant. Both the emitter-to-collector amplification (alpha) and the base-to-collector amplification (beta) can be checked in this manner.

Alpha measurement

The test circuit is shown in Fig. 908-a. The collector supply voltage (V_{cc}) and emitter current (i_e) levels at which the test will be made are chosen from the transistor characteristics sheet. In the test, i_e will be varied a small amount above and below the selected operating value, and the corresponding i_c change noted.



Figs. 908-a, -b. *Test circuit for current amplification, alpha (a). Test circuit for current amplification, beta (b).*

By adjustment of the output of the constant-current supply, set i_e to a value 0.1 ma lower than the selected emitter-current value. Record this setting as i_e1 and the corresponding collector current as i_c1. Next, set the emitter current to a point 0.1 ma higher than the selected value i_e. Record this new value as i_e2 and the corresponding collector current as i_c2. Alpha then is determined from the ratio of the collector and emitter current changes. Thus: $\alpha = (i_c2 - i_c1) / (i_e2 - i_e1)$.

Alpha measurements may be made at a variety of i_e and V_{cc} values to obtain complete data showing the variation of current amplification with these parameters.

Beta measurement

The test circuit for this measurement is shown in Fig. 908-b.

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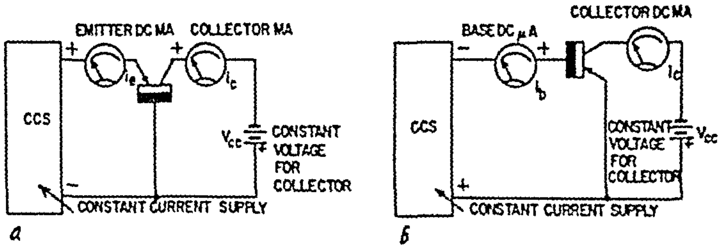

Figs. 908-a, -b. *Test circuit for current amplification, alpha (a). Test circuit for current amplification, beta (b).*

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Alpha measurements may be made at a variety of i_e and V_{cc} values to obtain complete data showing the variation of current amplification with these parameters.

Beta measurement

The test circuit for this measurement is shown in Fig. 908-b.

Voltage gain

Voltage gain, or *voltage amplification* may be checked with on of the setups shown in Fig. 909. The grounded-base circuit (Fi 909-a) may be used with point-contact and junction transistor The grounded-emitter (Fig. 909-b) and grounded-collector (Fi 909-c) are for junction types.

In each circuit, the transistor input electrode dc biases ar furnished by constant-current supplies. The output electrode bi is furnished by a constant-voltage supply. The sine-wave ac te signal is supplied through the input transformer T. The in pedance ratio of this component is not important, since onl a small input signal voltage is required at the transistor and co siderable mismatch loss can be tolerated. The input signal, applie to the transformer primary, is furnished by an oscillator or sign generator. A test frequency of 1,000 cycles is usual for lowe frequency measurements.

The test procedure is the same for each circuit: (1) Select value for the load resistance, R_L. (2) Adjust the transistor dire currents and voltages to their proper operating values accordin to the characteristics sheet for the type under test. The setting must be made with the ac signal absent. (3) Switch the ac vacuur tube voltmeter to the transistor input by throwing switch S t position 1. (4) By means of the output control in the signal ger erator, set the ac signal voltage to a convenient low value (1 t 10 millivolts rms), as indicated by the vacuum-tube voltmete Record this value as E_1. (5) Switch the vacuum-tube voltmete to the transistor output by throwing switch S to position 2, an read the signal output voltage. Record this value as E_2. (6) De termine the voltage gain from E_2/E_1.

The test may be repeated at any desired combination of tran sistor direct-current levels and at various values of R_L. An oscill scope should be operated across the load resistor, R_L to monito the output waveform. When severe distortion, such as peak clip ping, is evidenced, the input signal amplitude should be reduce

Input impedance

The input impedance of a transistor amplifier or ac contr circuit may be measured with the test setup shown in Fig. 91

The test signal is supplied by a suitable source, such as an osci lator or signal generator, having low-impedance output and co tinuously variable output control. A calibrated, noninductive variable resistor (rheostat or decade box) is connected in serie

Voltage gain

Voltage gain, or *voltage amplification* may be checked with one of the setups shown in Fig. 909. The grounded-base circuit (Fig. 909-a) may be used with point-contact and junction transistors. The grounded-emitter (Fig. 909-b) and grounded-collector (Fig. 909-c) are for junction types.

In each circuit, the transistor input electrode dc biases are furnished by constant-current supplies. The output electrode bias is furnished by a constant-voltage supply. The sine-wave ac test signal is supplied through the input transformer T. The impedance ratio of this component is not important, since only a small input signal voltage is required at the transistor and considerable mismatch loss can be tolerated. The input signal, applied to the transformer primary, is furnished by an oscillator or signal generator. A test frequency of 1,000 cycles is usual for lower-frequency measurements.

The test procedure is the same for each circuit: (1) Select a value for the load resistance, R_L. (2) Adjust the transistor direct currents and voltages to their proper operating values according to the characteristics sheet for the type under test. The settings must be made with the ac signal absent. (3) Switch the ac vacuum-tube voltmeter to the transistor input by throwing switch S to position 1. (4) By means of the output control in the signal generator, set the ac signal voltage to a convenient low value (1 to 10 millivolts rms), as indicated by the vacuum-tube voltmeter. Record this value as E_1. (5) Switch the vacuum-tube voltmeter to the transistor output by throwing switch S to position 2, and read the signal output voltage. Record this value as E_2. (6) Determine the voltage gain from E_2/E_1.

The test may be repeated at any desired combination of transistor direct-current levels and at various values of R_L. An oscilloscope should be operated across the load resistor, R_L to monitor the output waveform. When severe distortion, such as peak clipping, is evidenced, the input signal amplitude should be reduced.

Input impedance

The input impedance of a transistor amplifier or ac control circuit may be measured with the test setup shown in Fig. 910.

The test signal is supplied by a suitable source, such as an oscillator or signal generator, having low-impedance output and continuously variable output control. A calibrated, noninductive, variable resistor (rheostat or decade box) is connected in series

variable resistor. If the resistor is not calibrated for direct readings in ohms, it may be removed from the circuit and its resistance setting checked with a bridge or accurate ohmmeter.

For complete data, the input impedance should be checked at several values of R_L unless the latter has been established at a known fixed value. It should be checked also at various transistor bias levels.

Output impedance

The output impedance of a transistor amplifier or oscillator may be checked by means of an external resistor and ac vacuum-tube voltmeter. An amplifier must be driven by a signal source, as shown in Fig. 911, but an oscillator will supply its own output voltage for the test. In the case of an amplifier, the output must be monitored with an oscilloscope to prevent overdriving.

The variable resistor R must be noninductive. It may be a laboratory resistance decade box, although a composition potentiometer can be used. The test procedure is to measure the output voltage with no external load (switch S open), then to connect the load R (switch closed) and adjust R for a voltage reading equal to one-half the no-load voltage. At this point, the resistance setting of R equals the output impedance of the amplifier or oscillator. If the resistor is not calibrated to read ohms directly, its setting may be checked with a resistance bridge or accurate ohmmeter.

The dc output electrode current of the transistor must not be permitted to flow through the test circuit, otherwise the settings of

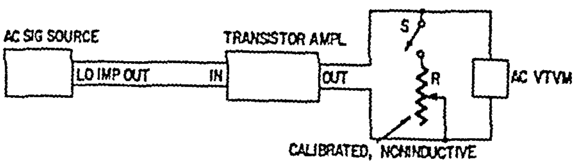

Fig. 911. *Circuit for measurement of output impedance.*

resistor R will shift the operating point of the transistor. The output impedance should be measured at various values of transistor dc electrode currents and, in the case of an amplifier, at various values of generator impedance.

Frequency response

Important transistor characteristics which vary with frequency are current amplification, voltage amplification, input impedance

current dc supply furnishes emitter bias, read with milliammeter M1, and the constant-voltage dc supply furnishes collector bias read with milliammeter M4. The large capacitors, C1 and C2, by-pass these supplies for ac.

The ac test signal is supplied through the secondary of transformer T. The test-signal amplitude should be between 0.1 and 0.5 volt rms across the secondary. A single transformer will not handle both audio and radio frequencies, and care must be taken when interchanging transformers and signal generators to keep

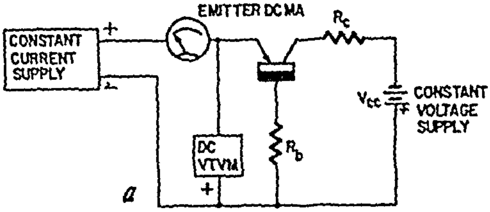

Fig. 913-a. *Circuit for testing emitter negative-resistance characteristics.*

the secondary impedance and dc resistance as nearly constant as possible.

The ac emitter current develops a voltage drop across the small series resistor R1. This drop is read as E1 with the first ac vacuum-tube voltmeter-millivoltmeter M2. Similarly, the ac col-

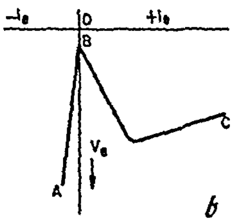

Fig. 913-b. *Negative-resistance characteristic curve.*

lector current develops a voltage drop E2 across R2, and this voltage is read with the second such meter, M3. Greatest convenience will be supplied by two separate meters. For economy, however, a switching arrangement may be used to switch a single meter between emitter and collector circuits. Alpha is the ratio of E2 to E1.

When making a run of alpha vs. frequency, voltage E1 must be maintained at a constant value. The output waveform should be

dc resistance of the transformer secondaries as nearly constant as possible. R_L is the rated load value.

Impedance vs.. frequency

Input and output impedance, as a function of frequency, can be checked with the circuits given in Figs. 910 and 911 by varying the test-signal frequency over the desired range. The test signal voltage and generator output impedance should be held constant. At each test frequency, the impedance should be checked in the manner described earlier.

Switching characteristics

The emitter and collector negative-resistance characteristics which suit point-contact transistors to switching circuits may be checked with the test circuits shown in Figs. 913 and 914, respectively. The values of R_b, R_e, and R_c, V_{cc}, and V_{ee} must be selected for the particular type of transistor under test.

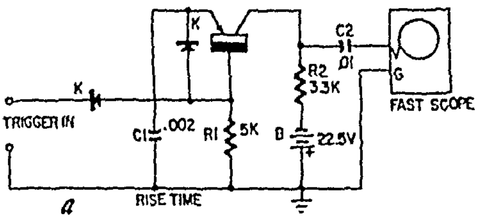

Fig. 915-a *Test circuit for measuring rise time of a switching transistor.*

Emitter negative resistance

In the test setup shown in Fig. 913-a, follow this procedure: (1) Reverse the polarities of the constant-current emitter dc supply and of the milliammeter temporarily. (2) Increase the emitter current in selected steps by adjusting the output of the constant-current supply. (3) Plot the corresponding values of negative emitter voltage to obtain the AB portion of the characteristic curve (Fig. 913-b). (4) Open the emitter connection temporarily to obtain the voltage reading at zero current. (5) Restore the connection and also the original polarity of the constant-current supply and milliammeter. (6) Increase the emitter current in steps and plot the corresponding emitter voltage points to obtain the remainder of the curve between B and C.

Collector negative resistance

Using the test circuit shown in Fig. 914-a, vary the collector

6. Rufus P. Turner, "Test Techniques for Transistors," RADIO-ELECTRONICS Magazine, March, 1953; p. 78.
7. L. J. Giacoletto, "Transistor Characteristics at Low and Medium Frequencies," *Tele-Tech*, March, 1953; p. 97.
8. Hans E. Hollmann, "Transistors in Terms of Vacuum Tubes," *Tele-Tech*, May, 1953; p. 74.
9. N. Golden and R. Nielsen, "Oscilloscopic Display of Transistor Static Electrical Characteristics," *Proceedings of the IRE*, November, 1952; p. 1437.
10. "A Simple Transistor Checker," *C-D Capacitor*, May, 1954.
11. Roland J. Turner, "Surface-Barrier Transistor Measurements and Applications," *Tele-Tech*, August, 1954; p. 79.
12. Rufus P. Turner, "Transistor Alpha-Beta Conversion," *Tele-Tech*, March, 1956; p. 74.
13. R. Zarr, "Transistor Tester," *Radio & Television News*, September, 1956; p. 140.
14. W. O. Hamlin, "Transistor Testing and Handling," RADIO-ELECTRONICS Magazine, February, 1957, p. 61.
15. Irving Shapiro, "Measurement Circuits for Sylvania Transistors," *Sylvania News*, January, 1958; p. 5.
16. "Simple Methods of Measuring Transistor Impedances," *C-D Capacitor*, February, 1958.
17. Roy A. Hempel, "A Transistor Tester for the Experimental Lab," *Electronic Industries*, February, 1958; p. 58.
18. John T. Frye, "Five New Transistor Checkers," RADIO-ELECTRONICS Magazine, March, 1958; p. 47.
19. Rufus P. Turner, "Transistor Alpha-Beta Nomogram," *Electrical Design News*, March, 1958; p. 50.
20. *Transistor Techniques*, Gernsback Library, Inc., Book No. 61; chapter 2, (pp. 17-32) and chapter 4 (pp. 45-56).

6. Rufus P. Turner, "Test Techniques for Transistors," RADIO-ELECTRONICS Magazine, March, 1953; p. 78.
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13. R. Zarr, "Transistor Tester," *Radio & Television News,* September, 1956; p. 140.
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germanium wafer in the region of the space charge and very near the p-type base region. This electrode is biased by voltage V_I so as to be slightly more negative than the potential gradient

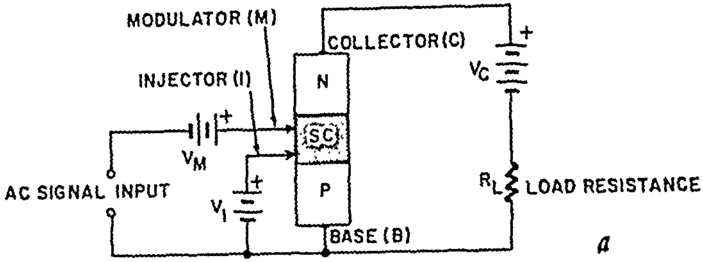

Fig. 1001-a. *Spacistor uses a reverse-biased p-n junction.*

at the point of contact. This causes electrons to be injected into the space-charge region SC and these carriers are accelerated rapidly in the electric field.

Another electrode, the *modulator* (M), is placed in contact

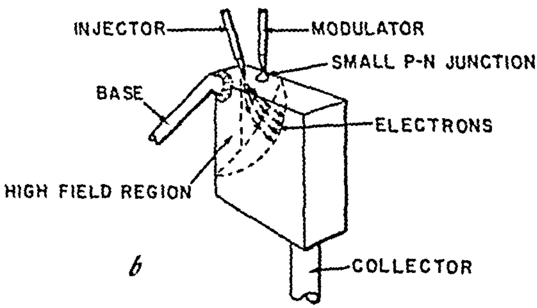

Fig. 1001-b. *Construction of the spacistor.*

with the space-charge region of the wafer between the injector and the n-type collector region. This electrode is doped with a p-type impurity material to form a p-n junction with the semiconductor body. Voltage V_M biases the modulator more negative than the potential gradient along the semiconductor body at its point of contact. No modulator current flows since the modulator p-n junction is reverse-biased.

An ac input signal is applied between the modulator and base, superimposed upon the dc modulator bias, as shown in Fig. 1001-a. The resulting voltage fluctuations vary the injector current, modulating the electron flow through the space-charge region. The collector-to-base current flowing externally through the load

germanium wafer in the region of the space charge and very near the p-type base region. This electrode is biased by voltage V_I so as to be slightly more negative than the potential gradient

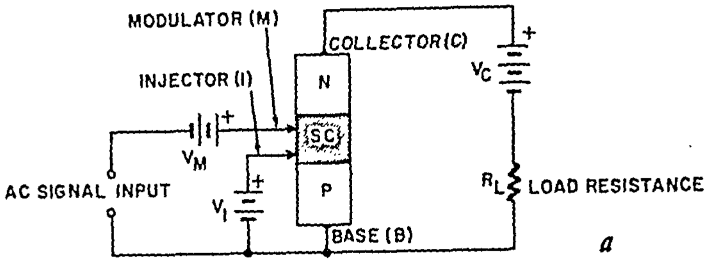

Fig. 1001-a. *Spacistor uses a reverse-biased p-n junction.*

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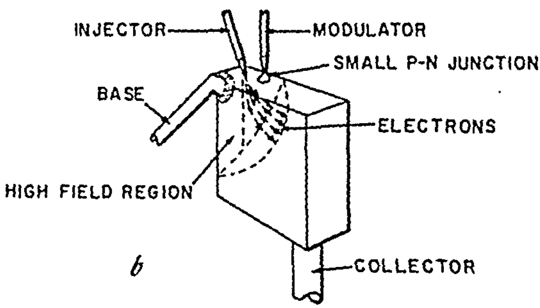

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of operation at 1000 mc and higher (with 6,000 mc as a probable upper limit). The output capacitance is reported as 1 μμf and lower. Fig. 1001-b shows the experimental construction of the spacistor.

Unijunction transistor

This semiconductor device originally was called a *double-base diode*. It utilizes the principle of injection and collection at a single junction.

The basic arrangement of the Unijunction is shown in Fig. 1002-a. A single p-n junction (emitter) is formed on an n-type silicon bar near its top end. An ohmic connection—*not a junction*—is made at each end of the bar. These are the two base connections of the device. A voltage source, V_{BB}, is conected between the bases.

When switch S is open, a voltage gradient along the bar is set up by V_{BB}. This gradient is almost uniform (that is, nearly linear) because the bar is acting as a resistance. However, since the emitter is near the top end of the bar (base 2) or more than halfway up, the voltage opposite the emitter is more than one-half V_{BB}. When S is closed, the p-n junction is forward-biased by V_e and hole curent flows between emitter and base 1. This lowers the resistance of the bar between base 1 and the emitter and produces a nonuniform distribution of the gradient along the bar.

When S is closed and V_e and V_{BB} are in the proper ratio, emitter action takes place between the p-n junction and base 2. This simultaneous injection and collection of carriers at the same junction gives rise to a negative input resistance similar to that described earlier for point-contact transistors. Fig. 1002-b shows this negative resistance characteristic. As the emitter current is increased from zero, the emitter-to-base voltage increases from A to B. As the current is increased further, the voltage falls from B to C. And as the current is increased still further, the voltage rises again slightly from C to D. Thus, a negative resistance region (B-C) is bounded by positive resistance regions (A-B and C-D). The interval between the peak point (B) and zero point is called the *cutoff region*, the interval between B and the *valley point* (C) the *negative resistance region*, and the interval between C and D the *saturation region*. The reason for the action which yields this response characteristic is that holes flowing into the lower part of the bar cause the voltage gradient along the bar to be redistributed and this changes the polarity of the emitter-to-base bias. At low values of current, the junction is positively-biased and acts as a

of operation at 1000 mc and higher (with 6,000 mc as a probable upper limit). The output capacitance is reported as 1 μμf and lower. Fig. 1001-b shows the experimental construction of the spacistor.

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true emitter, but at higher current levels the junction is negatively biased (as a result of the redistribution of the voltage gradient between base 1 and base 2), and acts as a collector.

The regenerativeness of this characteristic suits the unijunction to many simple circuits. As in the point-contact transistor, the

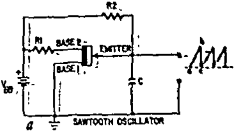


Fig. 1003-a. *Sawtooth oscillator using Unijunction transistor.*

negative-resistance region is unstable. However, it will provide monostable operation, as in one-shot multivibrator circuits, re-

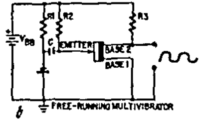


Fig. 1003-b. *Multivibrator using Unijunction transistor.*

laxation oscillators and similar circuits. By proper choice of [illegible] voltages and external circuit resistances, the load line (1-2-3 [illegible]

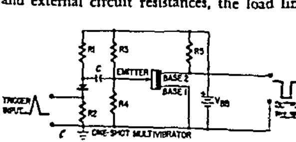


Fig. 1003-c. *Unijunction one-shot* [illegible]

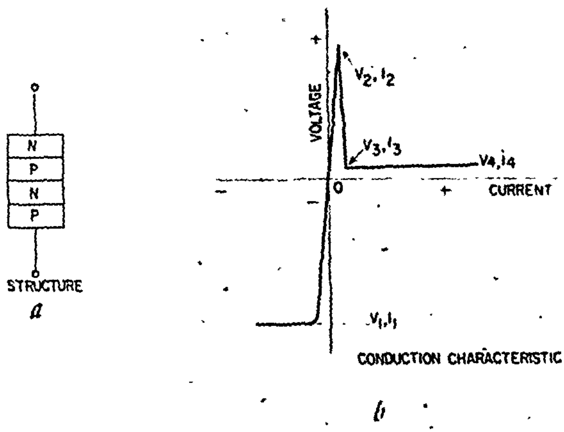
Fig. 1002-b) may be established and bi-stable [illegible]. Thus, with the Unijunction biased to the stable [illegible]

input pulse will shift the current through the unstable point 2 to the second stable point 3. The result is a simple flip-flop or scale-of-2 counter.

Fig. 1003 shows typical Unijunction circuits. A sawtooth oscillator is given in Fig. 1003-a. Here, the emitter-to-base voltage rises (from a to b in the accompanying waveform) while capacitor C is charging through resistor R2, and falls from b to c as C discharges. The frequency is governed by the R2C constants. If a suitable series resistor is inserted between base 1 and ground, steep positive pulses will be available at base 1 and similar negative pulses at base 2. Fig. 1003-b is a free-running multivibrator circuit. Fig. 1003-c shows a one-shot multivibrator.

Four-layer diode

The *four-layer diode* is a two-terminal device which has two stable states: high conductance (low resistance) and low conductance (high resistance). It exhibits effective switching action, since it is essentially closed in the first state and open in the second. Raising the applied voltage beyond a breakdown value causes the

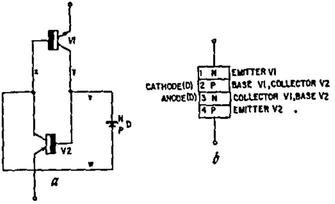


Figs. 1004-a, -b. *Structure and conduction characteristics of four-layer diodes.*

diode to flip to its closed state, and reducing the voltage below this level flips it back to the open state.

Fig. 1004-a shows the essential structure of this component. Alternate n and p layers are processed into the silicon body. Ohmic contacts are made to the respective n and p end regions. Fig. 1004-b

shows the conduction characteristic obtained with an a-c test signal. During the reverse half-cycle, voltage and current increase to V_1 and I_1, respectively. The current limiting at this point is due to



Figs 1005-a, -b. Equivalent circuit and structural arrangement of the four-layer diode.

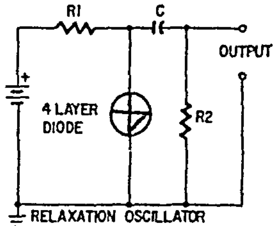
the avalanche[1] effect at voltage V_1. During the forward half-cycle, current and voltage increase to V_2 and I_2. A further increase in current to I_3 causes the voltage to decrease to V_3, indicating a negative resistance region. The forward characteristic of the four-layer voltage rises slowly to V_4, indicating saturation and a positive resistance region. As the current is increased still further to I_4, the voltage rises slowly to V_4, indicating saturation and a positive resistance region. The forward characteristic of the four-layer diode thus exhibits a negative resistance region similar to that of the point-contact and Unijunction transistors. Referring to Fig. 1004-b, V_2 is termed the *breakdown voltage*. When this voltage level is exceeded, the diode switches to its closed or ON state and will remain closed as long as I_3 (the *holding current*) is maintained. When the current is reduced below the level of I_3, the diode switches open or OFF.

The mechanism whereby the four-layer diode obtains [illegible]

[1]As the reverse voltage applied to certain semiconductor junctions is increased, the reverse current increases very slowly. At a certain point, however, the current increases very sharply as the voltage is further increased in small increments.

This condition of suddenly drawing large current resembles breakdown of the junction. However, it is not a true breakdown, since it is nondestructive if the maximum permissible power dissipation of the junction is not exceeded. Reducing the voltage below the "breakdown" point will restore the low-current condition. Nevertheless, the condition has been termed avalanche breakdown. Avalanche is noted particularly in the collector characteristic of a transistor and in the reverse conduction of diodes.

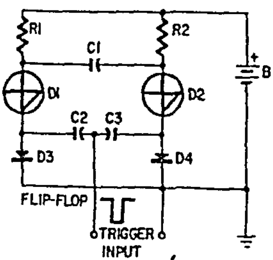
negative resistance and two stable states differs somewhat from other negative resistance devices. This component actually consists internally of a two-stage, direct-coupled, n-p-n–p-n-p, feedback transistor amplifier with an avalanche diode[2] connected between its collectors. Fig. 1005-a shows the equivalent circuit. The layers of the diode which provide the transistor and diode electrodes are identified in Fig. 1005-b. Thus, the n-p-n transistor (V1) is composed of layers 1, 2 and 3. The p-n-p transistor (V2) is



a

Fig. 1006-a. *Relaxation oscillator employing a four-layer diode.*

formed by layers 2, 3 and 4. In this way, the base of V1 is common with the collector of V2, and V1's collector is common with the



b

Fig. 1006-b. *Flip-flop circuit using two four-layer diodes.*

base of V2. This is equivalent to connections x and y in Fig. 1005-a. Also, the avalanche diode (D) is for[illegible] 2 and 3. But layer 2 is common to the base o[illegible] and this is equivalent to connection [illegible] the collector of V1 and base [illegible] connection v.

[2]A diode displaying the avalanche [illegible]

Fig. 1006-a is a relaxation oscillator employing a four-layer diode and having a sawtooth output voltage waveform. Fig. 1006-b shows a bi-stable (flip-flop) circuit employing two four-layer *diodes (D1 and D2) and two conventional diodes (D3 and D4).* Flip-flops of this type have been cascaded to form ring counters.

Solid-state thyratron

The silicon controlled rectifier behaves like a thyraton tube

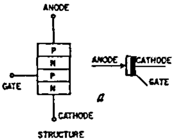


Fig. 1007-a. *Arrangement of the solid-state thyratron.*

in its ability to switch large amounts of current when turned on with a small control current and to remain on once triggered.

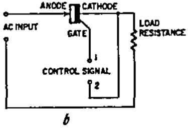


Fig. 1007-b. *Half-wave controlled-rectifier circuit using the solid-state thyratron.*

The structure of this device is represented by Fig. 1007-a. The four alternate n- and p-layers resemble the regions in the four-layer diode. The solid-state thyratron is indeed similar in that it consists essentially of two overlapping transistors, one p-n-p and one n-p-n (from top to bottom in Fig. 1007-a). The difference, however, is that a connection is made to the lower p-layer and this provides a *gate* electrode for triggering the device.

Fig. 1007-b shows the basic half-wave controlled-rectifier circuit

negative resistance and two stable states differs somewhat from other negative resistance devices. This component actually consists internally of a two-stage, direct-coupled, n-p-n–p-n-p, feedback transistor amplifier with an avalanche diode[2] connected between its collectors. Fig. 1005-a shows the equivalent circuit. The layers of the diode which provide the transistor and diode electrodes are identified in Fig. 1005-b. Thus, the n-p-n transistor (V1) is composed of layers 1, 2 and 3. The p-n-p transistor (V2) is

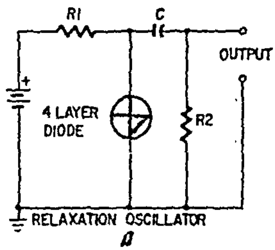


Fig. 1006-a. *Relaxation oscillator employing a four-layer diode.*

formed by layers 2, 3 and 4. In this way, the base of V1 is common with the collector of V2, and V1's collector is common with the

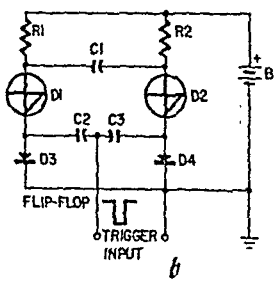


Fig. 1006-b. *Flip-flop circuit using two four-layer diodes.*

base of V2. This is equivalent to connections x and y in Fig. 1005-a. Also, the avalanche diode (D) is formed by layers 2 and 3. But layer 2 is common to the base of V1 and collector of V2 and this is equivalent to connection w. Also, layer 3 is common to the collector of V1 and base of V2, and this is equivalent to connection v.

[2]A diode displaying the avalanche characteristic is called an *avalanche diode.*

Fig. 1006-a is a relaxation oscillator employing a **four-layer** diode and having a sawtooth output voltage waveform. **Fig. 1006-b** shows a bi-stable (flip-flop) circuit employing two four-layer diodes (D1 and D2) and two conventional diodes (D3 and D4). Flip-flops of this type have been cascaded to form ring counters.

Solid-state thyratron

The silicon controlled rectifier behaves like a thyratron tube

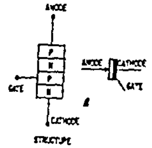


Fig. 1007-a. Arrangement of the solid-state thyratron.

in its ability to switch large amounts of current when turned on with a small control current and to remain on once triggered.

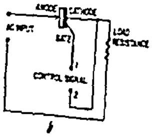
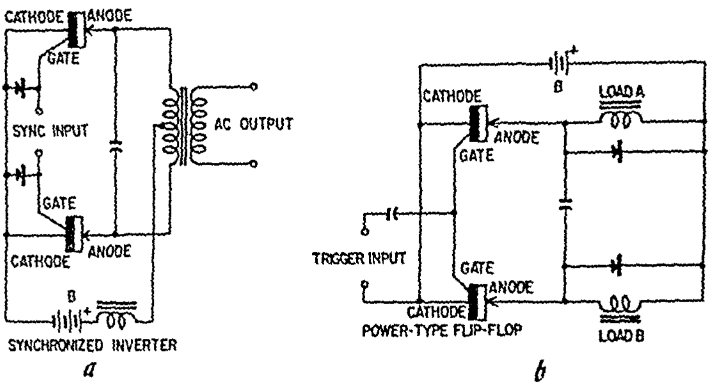


Fig. 1007-b. Half-wave controlled-rectifier circuit using the solid-state thyratron.

The structure of this device is represented by Fig. 1007-a. The four alternate n- and p-layers resemble the regions in the four-layer diode. The solid-state thyratron is indeed similar in that it consists essentially of two overlapping transistors, one p-n-p and one n-p-n (from top to bottom in Fig. 1007-a). The difference, however, is that a connection is made to the lower p-layer and this provides a *gate electrode* for triggering the device.

Fig. 1007-b shows the basic half wave controlled rectifier circuit

employing the solid-state thyratron. The rectifier blocks current during both the positive and negative half-cycles of ac input voltage unless the forward breakdown voltage is exceeded. When a positive gate-cathode current control pulse is applied to the gate electrode (input terminals 1 and 2), the device is "fired" by a reversal of bias on the center p-n junction, which causes the rectified current to increase sharply. This switches the device from nonconduction to full conduction. Extremely high-power control ratios (in the millions) have been obtained, and firing speeds in the vicinity of 1 microsecond noted.



Figs. 1008-a, -b. *Two applications of the solid-state thyratron. Synchronized inverter (a) and power type flip-flop (b).*

Fig. 1008 shows two applications of the solid-state thyratron. The inverter (Fig. 1008-a) may be synchronized (triggered) from a sine wave, pulses or square wave. Each time a trigger pulse is applied to the INPUT terminals of the power flip-flop circuit of Fig. 1008-b, the dc voltage is switched from one load to the other. The diodes are required only for inductive loads.

In addition to the typical examples shown, the solid-state thyratron has been applied to phase-controlled rectifiers, static switch circuits, dc-to-dc inverters, surge-voltage suppressor and dc-to-ac inverters. It has been employed also as a relay substitute and in dynamic braking. Early production models have controlled currents up to 5 amperes at 300 volts.

Phototransistor

In Chapter 1, it was mentioned that light energy will release current carriers in semiconductors. The resulting current is

but can be used, directly or indirectly, for many practical
oses An example of a photoelectric semiconductor device
e photocell employed in photographic exposure meters.
her is the germanium photodiode.

ne *phototransistor* exploits the principles of both photo
sion and current amplification. This device has the same
construction as the conventional transistor, except that a
low or lens is provided for light rays to impinge upon the

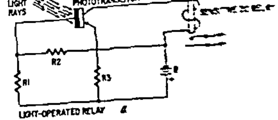


Fig 1009-a. *Light operated relay circuit employing a phototransistor*

miconductor surface. The circuit (Fig. 1009 ... is a
mmon-emitter amplifier. Here, however, the base-input signal

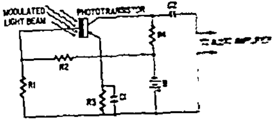


Fig. 1009-b *Phototransistor circuit using a modulated light beam.*

is not supplied by the usual external electrical source but consists of current flow resulting from the photoemission of carriers. This emission current is amplified in the usual manner by the transistor structure.

Fig. 1009-a shows a typical light-operated relay circuit employing a phototransistor. The dc operating point of the unit is stabilized by the base voltage divider (R1-R2) and the emitter current-limiting resistor (R3). Fig. 1009-b shows a simple phototransistor circuit for use with a modulated light beam. Applications include light-beam telegraphy, telephony and talking

pictures. A respectable amount of voltage amplification is provided by this arrangement.

Phototransistors are subminiature components. Their practical applications include punched tape and curve-trace reading in electronic computer applications; counting, sorting and color matching in industrial electronics, and a wide variety of control uses in household and commercial electronics.

Recommended Reading

1. J. J. Suran, "Double Base Expands Diode Applications," *Electronics,* March, 1955.
2. J. J. Suran, "Low-Frequency Circuit Theory of Double-Base Diode," *IRE Transactions of the Professional Group on Electronic Devices,* April, 1955.
3. A. William Carson, "Frequency Division with Semiconductor Devices," *Electronic Design,* July 15, 1957; p. 34.
4. Louis E. Garner, Jr., "Using the Unijunction," RADIO-ELECTRONICS Magazine, July 1957; p. 91.
5. "Ten Kmc Semiconductor Amplifier," *Electronic Equipment,* August, 1957; p. 6.
6. Dr. William Shockley, "Unique Properties of the Four-Layer Diode," *Electronic Industries & Tele-Tech,* August, 1957; p. 58.
7. Eric Leslie, "Two New Approaches to Amplification," RADIO-ELECTRONICS Magazine, November, 1957; p. 32.
8. T. P. Sylvan, "Determining Frequency of Unijunction Transistor Relaxation Oscillators," *Electronic Equipment Engineering,* March, 1958; p. 4.
9. R. P. Frenzel and F. W. Gutzwiller, "Solid-State Thyratron Switches Kilowatts," *Electronics,* March 28, 1958; p. 52.
10. Loy E. Barton, "The Thyristor," *Electronic Design,* March 19, 1958; p. 48.

care and handling of transistors

TRANSISTORS are capable of efficient operation and long life. They are also remarkably durable. In spite of these favorable properties, however, they can be damaged and therefore should not be handled, stored, installed or operated carelessly. Because they are semiconductor devices and of subminiature construction, transistors are sensitive to some environmental and operating conditions relatively unimportant to vacuum tubes.

Incorrect handling, installation or operation of transistors can lead to limited life and substandard performance. In the extreme case, the electrical characteristics of the transistor may be degraded permanently or the unit destroyed. The need for special care arises from the following factors: (1) Semiconductors are temperature-sensitive. (2) These materials are somewhat intolerant of current and voltage overloads. (3) Conduction areas and electrode spacings are small in transistors. (4) Lead separations likewise are small. (5) Internal connections often are made with extremely fine wires.

This chapter discusses the correct techniques for handling transistors and lists the pitfalls which must be avoided. Persons who intend to work with transistors should familiarize themselves with the rules before handling these components.

Temperature considerations

Temperature is one of the most important factors to be considered. A rise in junction temperature increases the currents flowing through a transistor, and this produces further heating. The effect is cumulative and the resultant *runaway* of current

pictures. A respectable amount of voltage amplification is provided by this arrangement.

Phototransistors are subminiature components. Their practical applications include punched tape and curve-trace reading in electronic computer applications; counting, sorting and color matching in industrial electronics, and a wide variety of control uses in household and commercial electronics.

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Temperature considerations

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may destroy the transistor. In milder cases, a temperature rise boosts the currents less severely but enough to shift the operating point of the transistor and completely alter circuit performance.

Do not operate transistors at ambient or junction temperatures in excess of the manufacturer's ratings. Also, do not store transistors at temperatures in excess of those temperature ratings. Stabilize the bias direct currents and voltages so that these values becomes less dependent upon the transistor characteristics.

Transistor data tables show typical and absolute maximum ratings at 25°C. Derating factors must be applied for operation at other temperatures. Typical temperature limits are 85°C for germanium and 150°C for silicon transistors.

Do not operate transistors near hot tubes or resistors. Mount power transistors in suitable heat sinks to prevent overheating.

Installation

When soldering or welding to transistor leads, provide a satisfactory heat sink *on the transistor side of the joint.* A simple method is to hold the lead with flat-nose pliers. Continue to hold the lead until the joint has cooled completely.

When desirable, insert transistors into sockets. For this purpose, the leads should be clipped to approximately 1/4 inch. To prevent heat damage to the transistors, solder all circuit leads to the socket terminals before inserting the transistors.

To prevent whipping or swinging of a transistor when equipment must be subjected to vibration, mount the transistors in suitable clips to hold them steady. A number of such clips are commercially available and some provide heat-sink action as well as mechanical support. While transistors have excellent shock and vibration characteristics, they should not be subjected needlessly to mechanical abuse.

In handling transistors, do not bend or flex their leads any more than necessary. Irreparable damage is done when a lead is broken off flush with the seal.

Before installing a transistor into a circuit, observe the following points: (1) Type of transistor, n-p-n or p-n-p. (2) Check the power supply polarity to ascertain that is it correct for n-p-n or p-n-p, whichever applies. (3) Identify the transistor leads; interchanging the base, collector and emitter may ruin the transistor. (4) Identify the transistor model; the same type of case is used for many different types.

Before potting with other components, ascertain from the tran-

sistor manufacturer whether the potting temperature will be injurious.

Determine if the metal case or envelope of the transistor is connected to one of the electrodes. When there is such a connection, transistors must be mounted far enough apart to prevent contact between their cases. Also, such a connection, frequently found in power transistors which must be bolted to the chassis for heat-sink purposes, prevents the chassis grounding of other points of the circuit.

Environment

In addition to temperature, other environmental factors also are of concern to designers and users of transistor equipment. These include shock, vibration, humidity, light and strong electrical, magnetic and radioactive fields.

The transistor must not be subjected to excessive mechanical shock and vibration. When operation in such an environment is unavoidable, the transistor must be suitably shock-mounted.

Modern transistors are hermetically sealed. Nevertheless, they should not be operated or stored in a highly humid environment without observing the precautions for the high-humidity operation of other electronic components. Every effort must be made to keep the transistor dry around its lead seal. Dampness has been a continual bugbear to semiconductor devices.

Transistors in transparent or translucent capsules are apt to be light-sensitive. Shield them from strong illumination. Protect a phototransistor from ambient light so that it will give full performance when activated by the control beam

Transistors should be protected from the effects of strong fields, electrostatic, magnetic and radioactive Do not mount transistors near strong permanent magnets or electromagnetic devices such as heavy transformers, chokes, solenoids, relays, loudspeakers, etc. The fine wire leads inside high-frequency transistors have been known to burn out from the effects of a strong rf field from a nearby transmitter.

Operation

Check a transistor completely before installation When a complete test is not possible, check i_{co} and collector current at several i_e levels. Also, check alpha or beta.

Operate a transistor safely within its current, voltage and power dissipation ratings. When operating at other than 25°C, apply the derating factors recommended by the manufacturer.

Do not subject the transistor to large transients which momentarily may exceed its ratings. A circuit which appears satisfactory in all other respects should be examined for switching transients. When checking new transistor circuits, advance the electrode voltages slowly to their final values. Switching them on suddenly at full value may produce harmful switching transients. It is best to switch off the power before inserting or removing a transistor from equipment. However, when this is not feasible, the common electrode connection (base in *common-base* circuit, emitter in *common-emitter* circuit or collector in *common-collector* circuit) should be made first and broken last.

Mount a power transistor in a satisfactory heat sink. This is especially important when the transistor is operated at its maximum power capability. Whenever practicable, include automatic compensator devices (such as thermistors, semiconductor diodes, voltage-divider networks or compensating resistors) to stabilize the power transistor biases against variation with temperature.

Recommended Reading

1. Rufus P. Turner, "Care of Transistors," *Radio & Television News*, March, 1953; p. 40.
2. Elbert Robberson, "Make Ready for Transistors," *Radio & Television News*, April, 1955; p. 40.
3. W. O. Hamlin, "Transistor Testing and Handling," RADIO-ELECTRONICS Magazine, February, 1957; p. 61.
4. Paul Penfield, Jr., "Transistor Heat Sinks," *Radio & TV News*, March, 1958; p. 70.
5. *Transistor Techniques*, Gernsback Library, Inc., New York, N. Y.; pp. 9–16.

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Tables and Data

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Useful Information

To find the circumference of a circle, multiply the diameter by 3.1416.

To find the diameter of a circle, multiply the circumference by .31831.

To find the area of a circle, multiply the square of the diameter by .7854

The radius of a circle × 6 283185 = the circumference.

The square of the circumference of a circle × .07958 = the area.

Half the circumference of a circle × half its diameter = the area.

The circumference of a circle × .159155 = the radius.

The square root of the area of a circle × .56419 = the radius.

The square root of the area of a circle × 1.12838 = the diameter.

To find the diameter of a circle equal in area to a given square, multiply a side of the square by 1.12838.

To find the side of a square equal in area to a given circle, multiply the diameter by .8862.

To find the side of a square inscribed in a circle, multiply the diameter by .7071.

To find the side of a hexagon inscribed in a circle, multiply the diameter of the circle by .500.

To find the diameter of a circle inscribed in a hexagon, multiply a side of the hexagon by 1.7321.

To find the side of an equilateral triangle inscribed in a circle, multiply the diameter of the circle by .866.

To find the diameter of a circle inscribed in an equilateral triangle, multiply a side of the triangle by .57735.

To find the area of the surface of a ball (sphere), multiply the square of the diameter by 3.1416.

To find the volume of a ball (sphere), multiply the cube of the diameter by .5236.

Doubling the diameter of a pipe increases its capacity four times.

To find the pressure in pounds per square inch at the base of a column of water, multiply the height of the column in feet by .433.

A gallon of water (U. S. Standard) weighs 8.336 pounds and contains 231 cubic inches. A cubic foot of water contains 7½ gallons, 1728 cubic inches, and weighs 62.425 pounds at a temperature of about 39° F.
These weights change slightly above and below this temperature.

In accordance with the standard practice approved by the American Standards Association, the ratio 25.4 mm = 1 inch is used for converting millimeters to inches. This factor varies only two millionths of an inch from the more exact factor 25.40005 mm, a difference so small as to be negligible for industrial length measurements.

Metric Measures

The metric unit of length is the meter = 39.37 inches.
The metric unit of weight is the gram = 15.432 grains.
The following prefixes are used for sub-divisions and multiples: Milli = $\frac{1}{1000}$, Centi = $\frac{1}{100}$, Deci = $\frac{1}{10}$, Deca = 10, Hecto = 100, Kilo = 1000, Myria = 10,000.

Metric and English Equivalent Measures

MEASURES OF LENGTH

Metric		*English*
1 meter	=	39.37 inches, or 3.28083 feet, or 1.09361 yards
.3048 meter	=	1 foot
1 centimeter	=	.3937 inch
2.54 centimeters	=	1 inch
1 millimeter	=	.03937 inch, or nearly 1-25 inch
25.4 millimeters	=	1 inch
1 kilometer	=	1093.61 yards, or 0.62137 mile

MEASURES OF WEIGHT

Metric		*English*
1 gram	=	15.432 grains
.0648 gram	=	1 grain
28.35 grams	=	1 ounce avoirdupois
1 kilogram	=	2.2046 pounds
4536 kilogram	=	1 pound
1 metric ton 1000 kilograms	=	.9842 ton of 2240 pounds 19.68 cwt. 2204.6 pounds
1.016 metric tons 1016 kilograms	=	1 ton of 2240 pounds

MEASURES OF CAPACITY

Metric		*English*
1 liter (= 1 cubic decimeter)	=	61.023 cubic inches .03531 cubic foot .2642 gal. (American) 2.202 lbs. of water at 62° F.
28.317 liters	=	1 cubic foot
3.785 liters	=	1 gallon (American)
4.543 liters	=	1 gallon (Imperial)

English Conversion Table

Length			
Inches	×	.0833	= feet
Inches	×	.02778	= yards
Inches	×	.00001578	= miles
Feet	×	.3333	= yards
Feet	×	.0001894	= miles
Yards	×	36.00	= inches
Yards	×	3.00	= feet
Yards	×	.0005681	= miles
Miles	×	63360 00	= inches
Miles	×	5280 00	= feet
Miles	×	1760 00	= yards
Circumference of circle	×	.3183	= diameter
Diameter of circle	×	3.1416	= circumference
Area			
Square inches	×	.00694	= square feet
Square inches	×	0007716	= square yards
Square feet	×	144.00	= square inches
Square feet	×	.11111	= square yards
Square yards	×	1296.00	= square inches
Square yards	×	9.00	= square feet
Dia. of circle squared	×	.7854	= area
Dia. of sphere squared	×	3 1416	= surface
Volume			
Cubic inches	×	.0005787	= cubic feet
Cubic inches	×	00002143	= cubic yards
Cubic inches	×	004329	= U. S. gallons
Cubic feet	×	1728 00	= cubic inches
Cubic feet	×	.03704	= cubic yards
Cubic feet	×	7.4805	= U. S. gallons
Cubic yards	×	46656 00	= cubic inches
Cubic yards	×	27.00	= cubic feet
Dia. of sphere cubed	×	.5236	= volume
Weight			
Grains (avoirdupois)	×	.002286	= ounces
Ounces (avoirdupois)	×	0625	= pounds
Ounces (avoirdupois)	×	.00003125	= tons
Pounds (avoirdupois)	×	16 00	= ounces
Pounds (avoirdupois)	×	.01	= hundredweight
Pounds (avoirdupois)	×	.0005	= tons
Tons (avoirdupois)	×	32000 00	= ounces
Tons (avoirdupois)	×	2000.00	= pounds

English Conversion Table

Energy			
Horsepower	×	33000.	= ft.-lbs. per min.
B. t. u.	×	778.26	= ft.-lbs.
Ton of refrigeration	×	200.	= B. t. u. per min.
Pressure			
Lbs. per sq. in.	×	2.31	= ft. of water (60°F.)
Ft. of water (60°F.)	×	.433	= lbs. per sq. in.
Ins. of water (60°F.)	×	.0361	= lbs. per sq. in.
Lbs. per sq. in.	×	27.70	= ins. of water (60°F.)
Lbs. per sq. in.	×	2.041	= ins. of Hg. (60°F.)
Ins. of Hg (60°F.)	×	.490	= lbs. per sq. in.
Power			
Horsepower	×	746.	= watts
Watts	×	.001341	= horsepower
Horsepower	×	42.4	= B. t. u. per min.
Water Factors (at point of greatest density—39.2°F)			
Miners inch (of water)	×	8.976	= U. S. gals. per min.
Cubic inches (of water)	×	.57798	= ounces
Cubic inches (of water)	×	.036124	= pounds
Cubic inches (of water)	×	.004329	= U. S. gallons
Cubic inches (of water)	×	.003607	= English gallons
Cubic feet (of water)	×	62.425	= pounds
Cubic feet (of water)	×	.03121	= tons
Cubic feet (of water)	×	7.4805	= U. S. gallons
Cubic feet (of water)	×	6.232	= English gallons
Cubic foot of ice	×	57.2	= pounds
Ounces (of water)	×	1.73	= cubic inches
Pounds (of water)	×	26.68	= cubic inches
Pounds (of water)	×	.01602	= cubic feet
Pounds (of water)	×	.1198	= U. S. gallons
Pounds (of water)	×	.0998	= English gallons
Tons (of water)	×	32.04	= cubic feet
Tons (of water)	×	239.6	= U. S. gallons
Tons (of water)	×	199.6	= English gallons
U. S. gallons	×	231.00	= cubic inches
U. S. gallons	×	.13368	= cubic feet
U. S. gallons	×	8.345	= pounds
U. S. gallons	×	.8327	= English gallons
U. S. gallons	×	3.785	= liters
English gallons (Imperial)	×	277.41	= cubic inches
English gallons (Imperial)	×	.1605	= cubic feet
English gallons (Imperial)	×	10.02	= pounds
English gallons (Imperial)	×	1.201	= U. S. gallons
English gallons (Imperial)	×	4.546	= liters

Metric Conversion Table

Length			
Millimeters	×	.03937	= inches
Millimeters	÷	25.4	= inches
Centimeters	×	.3937	= inches
Centimeters	÷	2.54	= inches
Meters	×	39.37	= inches (Act. Cong.)
Meters	×	3.281	= feet
Meters	×	1.0936	= yards
Kilometers	×	.6214	= miles
Kilometers	÷	1.6093	= miles
Kilometers	×	3280.8	= feet
Area			
Sq. Millimeters	×	.00155	= sq. in.
Sq. Millimeters	÷	645.2	= sq. in.
Sq. Centimeters	×	.155	= sq. in.
Sq. Centimeters	÷	6.452	= sq. in.
Sq. Meters	×	10.764	= sq. ft.
Sq. Kilometers	×	247.1	= acres
Hectares	×	2.471	= acres
Volume			
Cu. Centimeters	÷	16.387	= cu. in.
Cu. Centimeters	÷	3.69	= fl. drs. (U.S.P.)
Cu. Centimeters	÷	29.57	= fl. oz. (U.S.P.)
Cu. Meters	×	35.314	= cu. ft.
Cu. Meters	×	1.308	= cu. yards
Cu. Meters	×	264.2	= gals. (231 cu. in.)
Litres	×	61.023	= cu. in. (Act. Cong.)
Litres	×	33.82	= fl. oz. (U.S.P.)
Litres	×	.2642	= gals. (231 cu. in.)
Litres	÷	3.785	= gals. (231 cu. in.)
Litres	÷	28.317	= cu. ft.
Hectolitres	×	3.531	= cu. ft.
Hectolitres	×	2.838	= bu. (2150.42 cu. in.)
Hectolitres	×	.1308	= cu. yds.
Hectolitres	×	26.42	= gals. (231 cu. in.)
Weight			
Grams	×	15.432	= grains (Act. Cong.)
Grams	÷	981.	= dynes
Grams (water)	÷	29.57	= fl. oz.
Grams	÷	28.35	= oz. avoirdupois
Kilo-grams	×	2.2046	= lbs.

Metric Conversion Table (Cont.)

Weight			
Kilo-grams	×	35.27	= oz. avoirdupois
Kilo-grams	×	.0011023	= tons (2000 lbs.)
Tonneau (Metric ton)	×	1.1023	= tons (2000 lbs.)
Tonneau (Metric ton)	×	2204.6	= lbs.
Unit Weight			
Grams per cu. cent.	÷	27.68	= lbs. per cu. in.
Kilo per meter	×	.672	= lbs. per ft.
Kilo per cu. meter	×	.06243	= lbs. per cu. ft.
Kilo per Cheval	×	2.235	= lbs. per h. p.
Grams per liter	×	.06243	= lbs. per cu. ft.
Pressure			
Kilo-grams per sq. cm.	×	14.223	= lbs. per sq. in.
Kilo-grams per sq. cm.	×	32.843	= ft. of water (60°F.)
Atmospheres (international)	×	14.696	= lbs. per sq. in
Energy			
Joule	×	.7376	= ft. lbs.
Kilo-gram meters	×	7.233	= ft. lbs.
Power			
Cheval vapeur	×	.9863	= h. p.
Kilo-watts	×	1.341	= h. p.
Watts	÷	746.	= h. p.
Watts	×	.7373	= ft. lbs. per sec
Miscellaneous			
Kilogram calorie	×	3.968	= B. t. u.
Standard gravity (Sea level 45° lat.)	÷	980.665	= centimeters per sec. per sec.
Frigories/hr. (French)	÷	3023.9	= Tons refrigeration

The following pages show temperatures on Fahrenheit and Centigrade thermometers.

Equivalent Temperature Readings for Fahrenheit and Centigrade Scales

Fahrenheit Degs.	Centigrade Degs.	Fahrenheit Degs.	Centigrade Degs.	Fahrenheit Degs.	Centigrade Degs.	Fahrenheit Degs.	Centigrade Degs.
-459.4	-273	-21.	-29.4	17.6	-8.	56.	13.3
-436	-270.	-20.2	-29.	18.	-7.8	57.	13.9
-418.	-260.	-20.	-28.9	19.	-7.2	57.2	14.
-400.	-240	-19.	-28.3	19.4	-7.	58.	14.4
-382.	-230	-18.4	-28.	20.	-6.7	59.	15.
-364.	-220.	-18.	-27.8	21.	-6.1	60.	15.6
-346.	-210	-17.	-27.2	21.2	-6.	60.8	16.
-328.	-200	-16.6	-27.	22.	-5.6	61.	16.1
-310.	-190	-16.	-26.7	23.	-5.	62.	16.7
-292	-180	-15.	-26.1	24.	-4.4	62.6	17.
-274	-170	-14.8	-26.	24.8	-4.	63.	17.2
-256.	-160.	-14.	-25.6	25.	-3.9	64.	17.8
-238.	-150.	-13.	-25.	26.	-3.3	64.4	18.
-220.	-140	-12.	-24.4	26.6	-3.	65.	18.3
-202	-130.	-11.2	-24.	27.	-2.8	66.	18.9
-184	-120.	-11.	-23.9	28.	-2.2	66.2	19.
-166.	-110	-10.	-23.3	28.4	-2.	67.	19.4
-148.	-100	-9.4	-23.	29.	-1.7	68.	20.
-139.	-95.	-9.	-22.8	30.	-1.1	69.	20.6
-130	-90	-8.	-22.2	30.2	-1.	69.8	21.
-121.	-85.	-7.6	-22.	31.	-0.6	70.	21.1
-112	-80.	-7.	-21.7	32.	0.	71.	21.7
-103.	-75	-6.	-21.1	33.	+0.6	71.6	22.
-94.	-70	-5.8	-21.	33.8	1.	72.	22.2
-85.	-65	-5.	-20.6	34.	1.1	73.	22.8
-76	-60.	-4.	-20.	35.	1.7	73.4	23.
-67.	-55	-3.	-19.4	35.6	2.	74.	23.3
-58	-50	-2.2	-19.	36.	2.2	75.	23.9
-49.	-45	-2.	-18.9	37.	2.8	75.2	24.
-40.	-40.	-1.	-18.3	37.4	3.	76.	24.4
-39.	-39.4	-0.4	-18.	38.	3.3	77.	25.
-38.2	-39.	0.	-17.8	39.	3.9	78.	25.6
-38.	-38.9	+1.	-17.2	39.2	4.	78.8	26.
-37.	-38.3	1.4	-17.	40.	4.4	79.	26.1
-36.4	-38.	2.	-16.7	41.	5.	80.	26.7
-36.	-37.8	3.	-16.1	42.	5.6	80.6	27.
-35.	-37.2	3.2	-16.	42.8	6.	81.	27.2
-34.6	-37.	4.	-15.6	43.	6.1	82.	27.8
-34.	-36.7	5.	-15.	44.	6.7	82.4	28.
-33.	-36.1	6.	-14.4	44.6	7.	83.	28.3
-32.8	-36.	6.8	-14.	45.	7.2	84.	28.9
-32.	-35.6	7.	-13.9	46.	7.8	84.2	29.
-31.	-35.	8.	-13.3	46.4	8.	85.	29.4
-30.	-34.4	8.6	-13.	47.	8.3	86.	30.
-29.2	-34.	9.	-12.8	48.	8.9	87.	30.6
-29.	-33.9	10.	-12.2	48.2	9.	87.8	31.
-28.	-33.3	10.4	-12.	49.	9.4	88.	31.1
-27.4	-33.	11.	-11.7	50.	10.	89.	31.7
-27.	-32.8	12.	-11.1	51.	10.6	89.6	32.
-26.	-32.2	12.2	-11.	51.8	11.	90.	32.2
-25.6	-32.	13.	-10.6	52.	11.1	91.	32.8
-25.	-31.7	14.	-10.	53.	11.7	91.4	33.
-24.	-31.1	15.	-9.4	53.6	12.	92.	33.3
-23.8	-31.	15.8	-9.	54.	12.2	93.	33.9
-23.	-30.6	16.	-8.9	55.	12.8	93.2	34.
-22.	-30.	17.	-8.3	55.4	13.	94.	34.4

[illegible]

Equivalent Temperature Readings for Fahrenheit and Centigrade Scales

Fahren-heit Degs.	Centi-grade Degs.	Fahren-heit Degs.	Centi-grade Degs.	Fahren-heit Degs.	Centi-grade Degs.	Fahren-heit Degs.	Centi-grade Degs.
95.	35.	134.	56.7	172.4	78.	211.	99.4
96.	35.6	134.6	57.	173.	78.3	212.	100.
96.8	36.	135.	57.2	174.	78.9	213.	100.6
97.	36.1	136.	57.8	174.2	79.	213.8	101.
98.	36.7	136.4	58.	175.	79.4	214.	101.1
98.6	37.	137.	58.3	176.	80.	215.	101.7
99.	37.2	138.	58.9	177.	80.6	215.6	102.
100.	37.8	138.2	59.	177.8	81.	216.	102.2
100.4	38.	139.	59.4	178.	81.1	217.	102.8
101.	38.3	140.	60.	179.	81.7	217.4	103.
102.	38.9	141.	60.6	179.6	82.	218.	103.3
102.2	39.	141.8	61.	180.	82.2	219.	103.9
103.	39.4	142.	61.1	181.	82.8	219.2	104.
104.	40.	143.	61.7	181.4	83.	220.	104.4
105.	40.6	143.6	62.	182.	83.3	221.	105.
105.8	41.	144.	62.2	183.	83.9	222.	105.6
106.	41.1	145.	62.8	183.2	84.	222.8	106.
107.	41.7	145.4	63.	184.	84.4	223.	106.1
107.6	42.	146.	63.3	185.	85.	224.	106.7
108.	42.2	147.	63.9	186.	85.6	224.6	107.
109.	42 8	147.2	64.	186.8	86.	225.	107.2
109 4	43.	148.	64.4	187.	86.1	226.	107.8
110.	43.3	149.	65.	188.	86.7	226.4	108.
111.	43.9	150.	65.6	188.6	87.	227.	108.3
111.2	44.	150.8	66.	189.	87.2	228.	108.[illegible]
112.	44.4	151.	66.1	190.	87.8	228.2	109.
113.	45.	152.	66.7	190.4	88.	229.	109.[illegible]
114.	45.6	152.6	67.	191.	88.3	230.	110[illegible]
114 8	46.	153.	67.2	192.	88.9	231.	110[illegible]
115.	46.1	154.	67.8	192.2	89.	231.8	[illegible]
116.	46.7	154.4	68.	193.	89.4	232.	[illegible]
116 6	47.	155.	68.3	194.	90.	233.	[illegible]
117.	47.2	156.	68.9	195.	90.6	233.6	[illegible]
118.	47.8	156.2	69.	195.8	91.	234.	[illegible]
118.4	48.	157.	69.4	196.	91.1	235.	[illegible]
119.	48.3	158.	70.	197.	91.7	235.4	[illegible]
120.	48.9	159.	70.6	197.6	92.	236.	[illegible]
120.2	49.	159.8	71.	198.	92.2	237.	[illegible]
121.	49.4	160.	71.1	199.	92.8	237.2	[illegible]
122.	50.	161.	71.7	199.4	93.	238.	[illegible]
123.	50 6	161.6	72.	200.	93.3	239.	[illegible]
123.8	51.	162.	72.2	201.	93.9	240.	[illegible]
124.	51.1	163.	72.8	201.2	94.	240.8	[illegible]
125.	51.7	163.4	73.	202.	94.4	241.	[illegible]
125.6	52.	164.	73.3	203.	95.	242.	[illegible]
126.	52.2	165.	73.9	204.	95.6	242.6	[illegible]
127.	52.8	165.2	74.	204.8	96.	243.	[illegible]
127.4	53.	166.	74.4	205.	96.1	244.	[illegible]
128.	53.3	167.	75.	206.	96.7	244.4	[illegible]
129.	53.9	168.	75.6	206.6	97.	245.	[illegible]
129.2	54.	168.8	76.	207.	97.2	246.	[illegible]
130.	54.4	169.	76.1	208.	97.8	246.2	[illegible]
131.	55.	170.	76.7	208.4	98.	247.	[illegible]
132.	55.6	170.6	77.	209.	98.3	248.	[illegible]
132.8	56.	171.	77.2	210.	98.9	249.	[illegible]
133.	56.1	172.	77.8	210.2	99.	249.8	[illegible]